IRE Transactions

ON VEHICULAR COMMUNICATIONS

UNIVERSITY OF HAWAII LIBRARY
PERIODICAL
UNIVERSITY OF HAWAII LIBRARY

Volume VC-9 Number 1
Follows Vol. PGVC-13 MAY, 1960

TABLE OF CONTENTS

		Page
Impulse Noise Reduction Circuit for Communication Receivers	W. F. Chow	1
A New Concept in the Reliability Evaluation of Vehicular Radio	Frederick L. Hilton	10
A New Trend in Minified Communication Equipment	J. W. Knoll	25
Applications of Semiconductors in an 860 MC Receiver	L. G. Schimpf	33
A Hybrid Mobile Two-Way Radio	R. A. Beers, Jr.	38
Coordinated Broadband Mobile Telephone System	W. D. Lewis	43
Multi-Area Mobile Radio Telephone System	H. J. Schulte, Jr. and W. A. Cornell	49
Precision Carrier Frequency Control and Modulation Phase Equalization of Base Transmitters in a Mobile Radio System	D. S. Dewire and E. A. Steere	54
Comparison of Wideband and Narrowband Mobile Circuits at 150 MC	H. W. Nylund	69
A Transmission Line and Radiating System Measurement	W. F. Biggerstaff	75
An Omidirectional Medium Gain 460 MC Base Station Antenna Using Simplified Suppressors	J. S. Brown	80
Strictly Personal	George M. Dewire	86
Mobile Interference Symposium	J. F. Chappell, W. A. Shipman, K. E. H. Backman, and B. H. Short	91

TK7882
M6I13
VC-9

PUBLISHED BY THE
PROFESSIONAL GROUP ON VEHICULAR COMMUNICATIONS

IRE PROFESSIONAL GROUP ON VEHICULAR COMMUNICATIONS

The Professional Group on Vehicular Communications is an organization, within the framework of the IRE, of members with principal professional interest in Vehicular Communications. All members of the IRE are eligible for membership in the Group and will receive all Group publications upon payment of prescribed fees.

Annual Fee: $2.00

Administrative Committee

ANGUS A. MACDONALD, *Chairman*

R. P. GIFFORD, *Vice-Chairman*

J. J. RENNER, *Secretary*

S. C. BARTLETT, *Treasurer*

A. B. BUCHANAN
W. G. CHANEY
G. E. DODRILL
C. M. HEIDEN

Ex-Officio:

J. E. KELLER
M. E. KENNEDY
L. E. LUDEKENS
NEWTON MONK
M. E. FLOGEL
C. R. KRAUS

J. R. NEUBAUER
J. S. STOVER
W. H. WATKINS
W. J. WEISZ
E. L. WHITE
K. J. WORTHEN

IRE TRANSACTIONS®
on Vehicular Communications

Published by the Institute of Radio Engineers, Inc., for the Professional Group on Vehicular Communications, 1 East 79th Street, New York 21, New York. Responsibility for the contents rests upon the authors, and not upon the IRE, the Group, or its members. Individual copies available for sale to IRE-PGVC members at $1.30, to IRE members at $1.95 and to nonmembers at $3.90.

COPYRIGHT ©1960 – THE INSTLTUTE OF RADIO ENGINEERS, INC.

All rights, including translation, are reserved by the IRE. Requests for republication privileges should be addressed to the Institute of Radio Engineers, 1 East 79th Street, New York 21, N. Y.

PRINTED IN THE U.S.A.

IMPULSE NOISE REDUCTION CIRCUIT FOR COMMUNICATION RECEIVERS

W. F. Chow
General Electric Company
Syracuse, New York

Summary

In this paper a comparatively simple scheme of reducing the impulse noise is suggested. This system consists of a gating signal generator and a balanced gate. It functions as follows: the impulse noise in the antenna triggers the gating signal generator. This generator produces a gating signal which in turn controls a balance gate. The function of the gate is to block the transmission during the gating period. Both the principle and the circuits of this impulse noise reduction scheme are described. Experimental results give the improvement of signal to noise ratio at the audio output ranging from 14 to 18 db.

Introduction

The problem of impulse noise has long been experienced by people in the radio communication field, especially with mobile communication equipment. The type of noise referred to here consists of impulses randomly spaced in time and generally of random amplitude. Atmospheric and man-made disturbances are the primary source of this type of noise. Distinct among these is the impulse noise produced by the automobile ignition system. This type of noise is characteristically composed of pulses of very short duration and very fast rise time occurring at a nearly periodic rate which is dependent upon the engine speed.

When an impulse noise is induced in the antenna of a receiver, it produces ringing in the selective circuits as it progresses from stage to stage. The shape of the envelope of this ringing is a function of the shape of the selectivity curve. There is a tendency for the voltage envelope to rise to a peak, drop to zero, and then reappear in the form of a secondary lobe with the phase of the radio frequency wave reversed. The size of the secondary lobe is a function of the shape of the nose of the selectivity curve. Very often an impulse of duration of only a fraction of a microsecond at the antenna produces a ringing of more than 300 microseconds long in the latter part of the receiver after passing through several selective circuits. In addition, there is the effect of impulse amplitude limiting, where the peak amplitude of the interfering impulse is considerably greater than that of the desired signal. This will occur at any point in the receiving system where the amplitude of the noise impulse is greater than the dynamic signal handling capacity.

Both AM and FM interference are produced by the impulse noises. In FM receivers the alignments of the IF filter and the discriminator have been found to have an important effect on the actual audio noise output. Since there is unavoidable misalignment or detuning after a period of field use, the impulse noise problem may be very severe.

Different schemes of reducing the impulse noise have been tried by many engineers. The results were either not satisfactory or the circuits involved were too complicated for practical use.

In this paper a comparatively simple scheme is suggested. Its principle and experimental results are described.

Principles

The duration of an impulse noise itself is rather short but it produces ringing which gets longer as it progresses in the receiver from the antenna towards the audio output. It is desirable to remove the noise at as early a stage as possible. Theoretically, one would like to have it removed right at the antenna input point but practical consideration that a circuit requires a finite amount of time to react demands that a time delay be provided between the antenna and the receiver input terminals. This time delay requirement may not easily be incorporated in the equipment.

The noise reduction technique described in this paper utilizes several facts: 1) The induced impulse noise at the receiver antenna is of much larger amplitude than the desired signal. 2) The duration of this induced noise may be very short and it may have short rise and fall times. However, there are low frequency components far below the RF in the induced noise pulse. 3) The receiver is of narrow band type and hence there are several RF selective circuits in the front end. 4) The audio output of the receiver is also restricted to a narrow band of frequency.

Figure 1 shows the block diagram of an FM receiver using two conversion schemes. The relative time scale of the signal and the impulse noises at points (1) and (2) are also shown. The signal and the noise are delayed δ second going through the RF amplifiers and the mixer stage. The duration of the impulse noise is also broadened a little going through several selective networks in the front end. However, the pass band of the front end selective circuits is relatively broad, the duration of the impulse noise is still small (may be of the order of a microsecond. It is not critical in this new system).

At the output side of this receiver, the audio system has a band from f_1 to f_2 (such as

300 cps to 3000 cps). The attenuation of frequencies beyond f_2 is designed to be very high.

Figure 2 shows the combination of the receiver block diagram and the block diagrams of the impulse noise reduction system. This system consists of a gating signal generator and a balanced gate and functions as follows: the impulse noise in the antenna triggers the gating signal generator. This generator produces a gating signal at point (3) after a delay of ξ second. This gating signal is wider than the duration of the impulse noise at point (2) and the delay, ξ , is made smaller than δ so that the gate is closed soon enough and nothing goes to the IF amplifier during the gating period. Consequently at point (4) the signal appears interrupted as shown in Figure 3.

The question whether the gate is capable of opening and closing without producing transients of its own is a matter of proper design of the gate circuit. This subject will be treated later.

The sudden interruption of RF signal may produce two objectionable results: One is the transient of the RF signal. The amplitude of this transient depends upon the attenuation characteristic of the gate and the amplitude of the RF signal just prior to the gating. The other is the noise at the output due to the disappearance of RF while the squelch is open. In order to overcome this difficulty, it is suggested that the gating signal generator in this system, when triggered by an impulse noise in the antenna, not only produce one gating signal to gate out the noise but a chain of gating signals evenly spaced with respect to time as shown in Figure 4. Consequently at point (4) the RF signal is interrupted as many times as the number of gating signals. It can be shown that if the repetition rate in cycles per second of the gating signal between the two triggering impulse noises is larger and beyond the audio band of the receiver, the noise produced by the interruption of RF signal at point (4) cannot be heard and the desired information in the RF can be recovered without distortion.

It is obvious that the higher the repetition rate, the more information will be lost. Therefore, more receiver noise will be present in the output. On the other hand, the frequency spectrum of the noise will be outside the audio band of the receiver. Let the gating period be a μsecond and the impulse noise repetition rate 1/b, as shown in Figure 5. The gate signal generator repetition rate is 1/c, where $1/c > 1/b$. Let the receiver white noise be N and the total peak noise due to interruption of the RF signal be N' over the period of gating. The value N' is a function of the RF signal and becomes zero when the RF signal is not present and N' is much smaller than the impulse noise. The idealized audio response of the receiver is shown in Figure 6. There is a slope of attenuation of k_1 db per octave for frequencies below 300 cps and k_2 db per octave for frequencies beyond 3000 cps.

Case a

When there is no impulse noise, the gate signal generator is not active. Let the signal be S, the signal-to-noise power ratio is $\frac{S}{N} = A$.

Case b

When there is impulse noise present and the gate generator is designed to produce one gating signal per noise pulse, then the signal power is reduced to $S\left(1 - \frac{a}{b}\right)$ and the noise power output is increased slightly due to N'. Expanding this noise into a Fourier's series results in:

$$\sum_{n=1}^{\infty} 2N' \frac{a}{b} \left[\frac{\sin n\pi \frac{a}{b}}{n\pi \frac{a}{b}} \right] \cos\left(\frac{2\pi nt}{b}\right)$$

$$= \sum_{n=1}^{\infty} C_n \cos\left(\frac{2\pi nt}{b}\right)$$

The total noise output will be:

$$N + \left\{ \frac{1}{2} \sum_{n=1}^{\infty} \left[C_n \text{ Resp } (f) \right]^2 \right\}^{\frac{1}{2}}$$

where Resp (f) represents the frequency response of the audio circuit. The signal-to-noise power ratio is then:

$$B = \frac{S\left(1 - \frac{a}{b}\right)}{\left[N + \left\{ \frac{1}{2} \sum_{n=1}^{\infty} \left[C_n \text{ Resp } (f) \right]^2 \right\}^{\frac{1}{2}} \right]} \quad (1)$$

Case c

When there is impulse noise present and the gate signal generator is designed to produce a gating signal at a rate of $\frac{1}{c}$ cycles per sec, then the signal is reduced to $S\left(1 - \frac{a}{c}\right)$ and the noise output is increased due to N'. The noise is

$$\sum_{n=1}^{\infty} 2N' \frac{a}{c} \left[\frac{\sin n\pi \frac{a}{c}}{n\pi \frac{a}{c}} \right] \cos\left(\frac{2\pi nt}{c}\right)$$

$$= \sum_{n=1}^{\infty} C'_n \cos\left(\frac{2\pi nt}{c}\right)$$

Total noise output will be

$$N + \left\{ \frac{1}{2} \sum_{n=1}^{\infty} \left[C'_n \text{ Resp } (f) \right]^2 \right\}^{\frac{1}{2}}$$

The signal-to-noise power ratio is

$$C = \frac{S\left(1 - \frac{a}{c}\right)}{\left[N + \left\{ \frac{1}{2} \sum_{n=1}^{\infty} \left[C'_n \text{ Resp } (f) \right]^2 \right\}^{\frac{1}{2}} \right]} \quad (2)$$

An idealized example is given below in order to compare Cases b and c. The frequency response of the audio circuit is given in Figure 6. Let a = 10 μsec, $\frac{1}{b}$ = 50cps, k_1 = 12, k_2 = 36 and N' = 50 N

Then from Equation (1)

$$B \cong S(1-0.5\times10^{-3}) \Big/ N\left\{1+\frac{25\times10^{-3}}{\sqrt{2}}\left[\sum \text{Resp}(f)^2\right]^{\frac{1}{2}}\right\}$$

$$\cdot \frac{S}{N}\frac{(1-0.5\times10^{-3})}{1+0.136} = \frac{S}{N}\times\frac{0.9995}{1.136}$$

$$\cong 0.884\,\frac{S}{N}$$

The noise output of Case c depends upon the repetition rate of the gating signal. The values of C in Equation (2) are calculated for different repetition rates and the ratio $\frac{C}{B}$ is plotted in Figure 7. For this example the improvement of signal-to-noise ratio begins from a repetition rate slightly larger than 5 kc to about 11 kc. The amount of improvement is only about 33% for the present case of N' = 50 N, $\frac{1}{b}$ = 50, and the assumed audio response. The height of the shaded area under the curve is an indication of the amount of improvement. This value is a function N', rate of noise pulse and the audio response.

Experimental Study

Receiver Response Without Noise Reduction Circuit

A 40 mc FM communication receiver was tested before modification. The bandwidth is 15 kc and the sensitivity is about 0.5 μv for 20 db quieting.

The measured time delay of a pulse between the antenna terminal through RF filters and the plate of the first mixer is about 1.2 μsec. The shape of the envelope of the ringing at this point depends upon the noise pulse shape and amplitude. For a pulse of about 0.2 μsec width and a peak amplitude of about 10 mv, the observed envelope shape is approximately as shown in Figure 8. The envelopes of ringing at different points in the receiver are shown in Figure 9a to 9e. The time delay of the pulse becomes longer and longer as it progresses towards the limiter. It increases into the order of magnitude of millisecond.

The time delay (about 1.2 μsec) of the noise pulse at the plate of the first mixer is more than enough for the present gating application.

Property of Gate

The function of the gate is to block the transmission between the plate of the first mixer and the high IF filter when a noise pulse is induced at the antenna. Ideally the gate should have a negligible insertion loss during normal receiving conditions and should produce an infinite attenuation during the gating period. An additional important requirement is that the gate itself should produce no transient.

The diode gate tested is shown in Figure 10. Two germanium point contact diodes 1N191 are connected in series as shown. They are normally forward biased with about 1ma each. Two silicon junction diodes (1N461) are connected in shunt as shown. They are normally reverse biased by the voltage drop across the germanium diodes. During the gating period the gating signal induced in the pulse transformer reverses both the germanium diodes and the silicon junction diodes. Thus, during this period the two silicon junction diodes are conducting and the germanium point contact diodes are reverse biased by the voltage drop across the silicon diodes. The response of this gate is shown in Figure 11.

Video (or Pulse) Amplifier and Gate Signal Generator

Although the noise pulse induced at the antenna is much larger than the desired RF signal, it is often much too small to be used to directly trigger the gate signal generator. The amount of amplification required depends upon the trigger voltage required in relation to the available noise pulse at the antenna. The most efficient way of utilizing the amplification per stage is to use pulse transformers. The other way is to include a video amplifier. In either case, an important requirement is that a limiter circuit should be included into the amplifier so that the output of the amplifier should not be larger than a fixed amount. This is to insure that no stage of the amplifier will be saturated when the noise pulse at the antenna is large. The reasons for preventing saturation are that when the triggering voltage from the video or pulse amplifier is too large the output wave form of the gate signal generator may be altered and when one or more stages of the video or pulse amplifier are saturated, the desired triggering signal will be delayed. Consequently the output of the gate signal generator will be out of timing with the noise ringing in the receiver.

The video amplifier tested is shown in Figure 12. Note that there are pairs of diodes in the collector circuit of each amplifier stage. For the first and the second stage, germanium point contact diodes (1N191) are connected in parallel. The last stage has two silicon junction diodes (1N461) connected to ground. These diodes are used to limit the voltage swing of each stage to about ± 0.2 volt peak for the germanium diodes and about ± 0.7 volt peak for the silicon diodes. There is another germanium point contact diode in series with the output of the last stage poled in such a way that the triggering pulse output of the amplifier can only turn on the gate signal generator. This diode prevents any possible interruption of the gate signal generator during the gating period by additional noise pulses.

The frequency response of this video amplifier is shown in Figure 13. The voltage amplification is approximately 80 db at the collector of the last stage. A pulse amplifier tested is shown in Figure 14. The estimated power gain is about 60 db.

The gate signal generator is a blocking oscillator either of the one-shot type or the free-running type. The circuit for the one-shot type is shown in Figure 15a. The circuit for the free-running type is the same except for biasing the transistor in the proper direction as shown in Figure 15b. The gating signal wave form depends upon the circuit design and the pulse transformer. The circuit tested has a gating period approximately 10 μsec as shown in Figure 16.

Receiver Response With Noise Reduction Circuit

The three stage video amplifier, the one-shot blocking oscillator and the diode gate were connected together as shown in Figure 17. The required transistor power supply was obtained with a voltage dividing network from the available +180 V supply in the FM receiver. The diode gate was inserted between the plate of the first mixer and the high IF filter. The input of the video amplifier was connected to the antenna terminal of the receiver. A switch was connected in the power supply circuit of the video amplifier and the gate signal generator so that a comparison of receiver response could be made with and without the gate signal.

After modification the receiver sensitivity was not changed. However, there was about 5 to 6 db loss of gain due to the diode gate. With the power supply switch of the video amplifier and the gate signal generator open, the response of the receiver to an impulse noise induced at the antenna was the same as before. The envelope of the ringing at the plate of the first limiter is shown in Figure 18a. Figure 18b shows the wave form at the same point when the noise reduction circuit was turned on. The residue ringing was about twice the amplitude of the receiver white noise. With proper design this residue ringing can be reduced further. The sensitivity of the gate was about 100 μv at the antenna. The improvement of signal to noise ratio at the audio output was 14 db. This gate did not respond to any RF signal either FM or AM at any level from zero up to several volts at the antenna. Therefore, there is no danger of triggering the gate signal generator by the desired RF.

The improvement of noise output by using high-rate gating was also tested. The free-running gate was synchronized by the amplified impulse. The repetition rate of the gate signal generator was about 5.5 KC. There was a range of RF signal level about 10 μV to 40 μV within which the improvement was very noticeable. An improvement of about 4 db over and above the improvement made by the one shot gating was obtained. For an RF signal level between 5 - 10 μV, no improvement was noticed. When the RF signal level was below about 5 μV, high rate gating slightly deteriorated the signal-to-noise ratio. When the RF signal level was larger than 50 μV, no improvement was noticed.

Conclusion

Experimental results indicate that the noise reduction circuit described above worked very well. The sensitivity of the circuit is limited only by the gain of the pulse amplifier. The number of components needed can be greatly reduced by modifying the existing amplifier stages to serve multiple functions.

Acknowledgement

The author wishes to express his gratitude to Messrs. D. C. Pinkerton and N. H. Shepherd for their encouragement, and their interest in the subject and to Mr. F. V. Adamthwaite for his assistance in the field testing of the equipment.

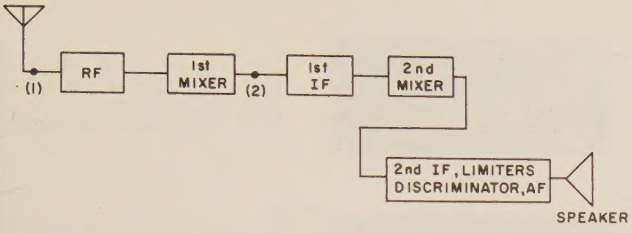
References

1. Lamb, J. J., "A Noise-Silencing IF Circuit for Superhet Receivers," QST P. 11, Feb. 1936.

2. Landen, V. D., "Impulse Noise in FM Reception," Electronics, Vol. 14, P. 26, Feb. 1941.

3. Wilmotte, R. M., "Reduction of Interference in FM Receiver by Feedback Across the Limiter," Proc. IRE, Vol. 40, PP. 34-36, Jan. 1952.

4. Lapin, S. P. and Suran, J. J., "Impulse Noise in Narrow-Band FM Receivers," Trans. AIEE, Vol. 74, Part I, PP. 450-454, Sept. 1955.



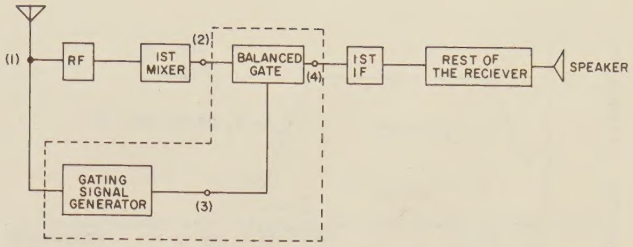


Fig. 2 - Block Diagram of the Impulse Noise Reduction System

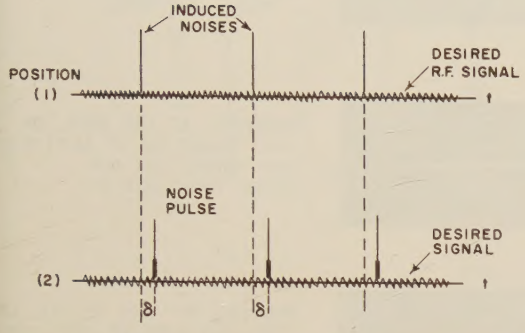
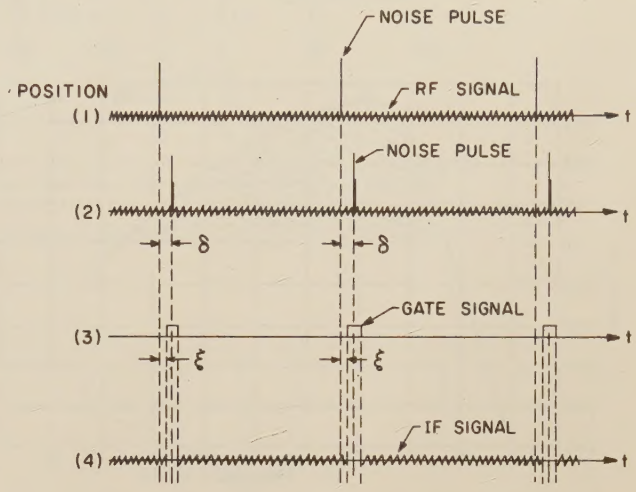


Fig. 1 - Block Diagram of Receiver and Delay of Noise Pulses

Fig. 3 - Relative Time of Noise and Gating Signal (Photo at right)



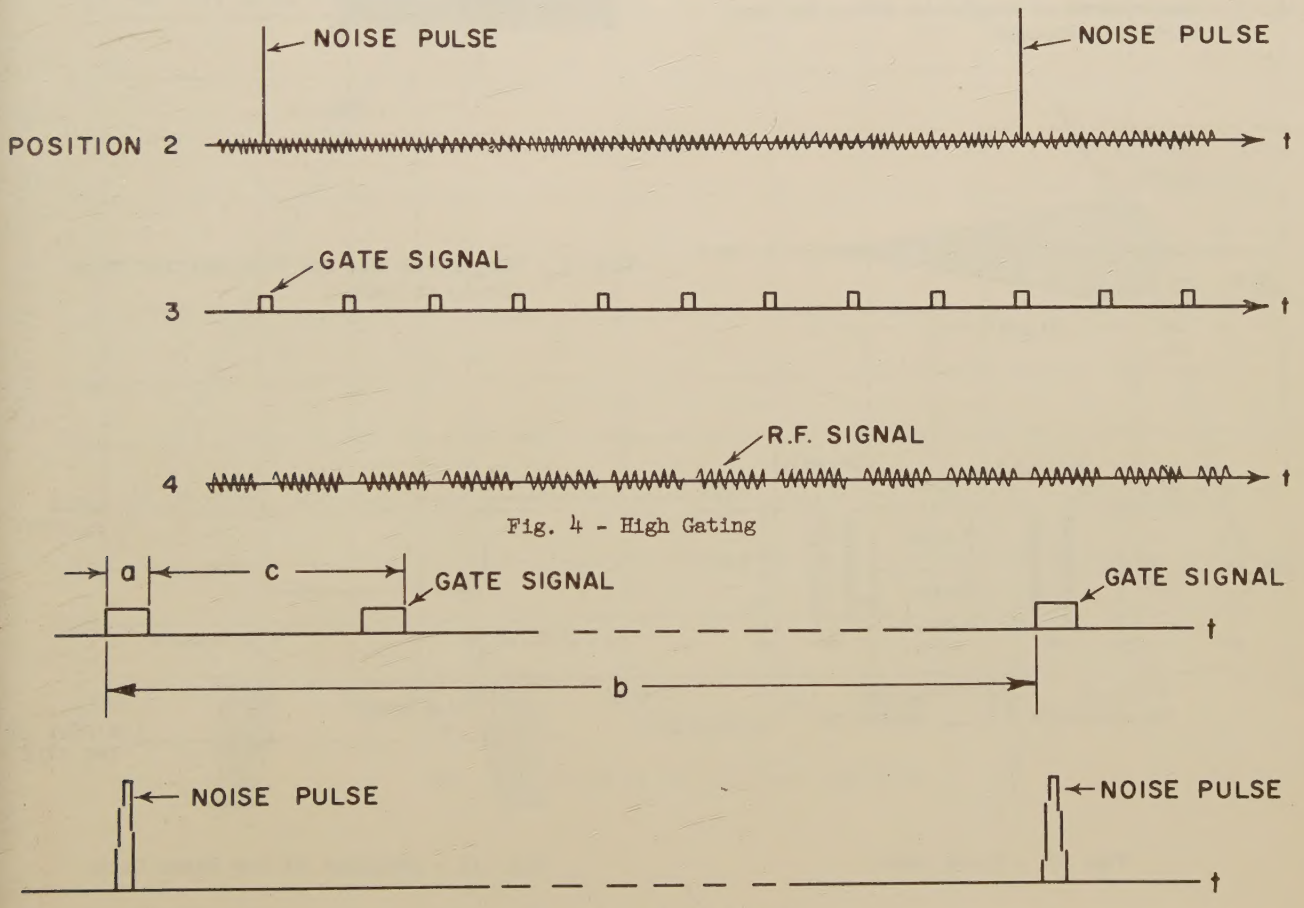


Fig. 4 - High Gating

Fig. 5 - Gating Period

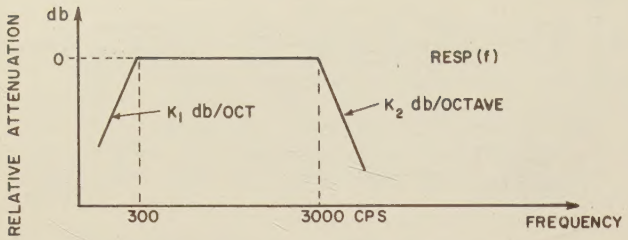


Fig. 6 - Idealized Audio Response

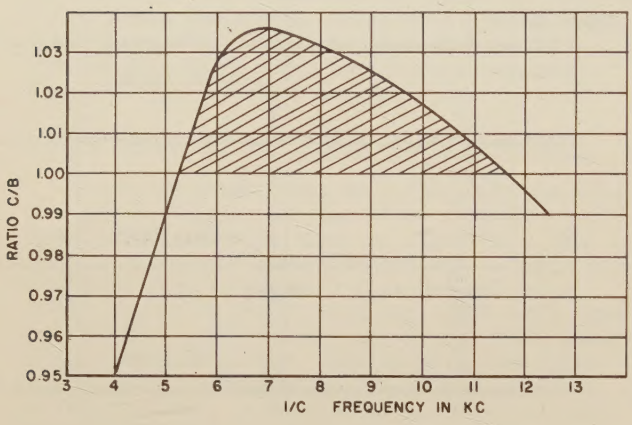
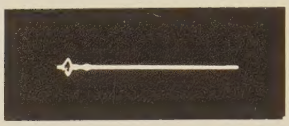


Fig. 7 - Improvement of Signal to Noise for the Idealized Example

a

RINGING AT THE GRID OF THE 1st LOW IF AMPLIFIER
MAX. PEAK 10 MV
100 μ SEC/MAJOR DIVISION

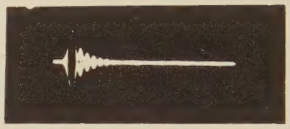


b

RINGING AT THE PLATE OF THE 1st LOW IF AMPLIFIER
MAX. PEAK 30 MV
SAME TIME SCALE

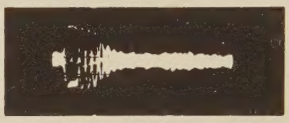
c

RINGING AT THE GRID OF THE 2nd LOW IF AMPLIFIER
MAX. PEAK 160 MV
SAME TIME SCALE



d

RINGING AT THE PLATE OF THE 2nd LOW IF AMPLIFIER
MAX. PEAK 9V
SAME TIME SCALE



e

RINGING AT THE PLATE OF THE 1st LIMITER
MAX. PEAK 110 V
SAME TIME SCALE

Fig. 9

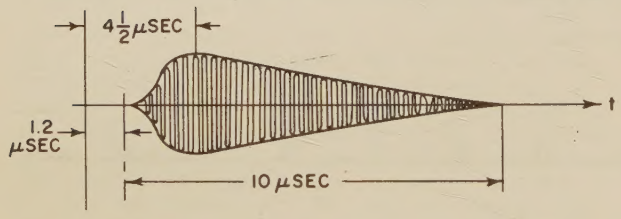


Fig. 8 - Ringing at the Plate of the 1st Mixer (Photo at left)

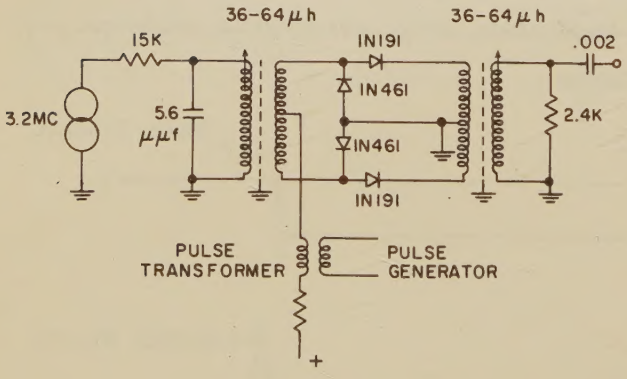


Fig. 10 - Diode Gate

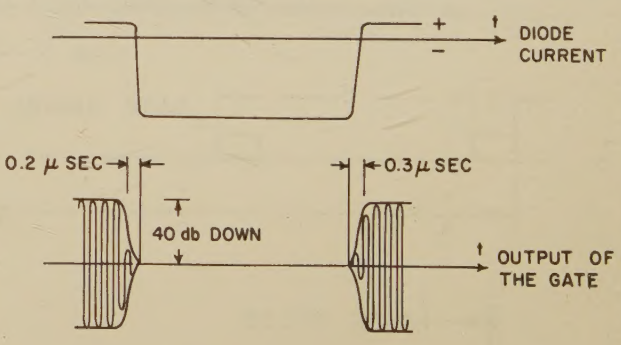


Fig. 11 - Response of the Diode Gate

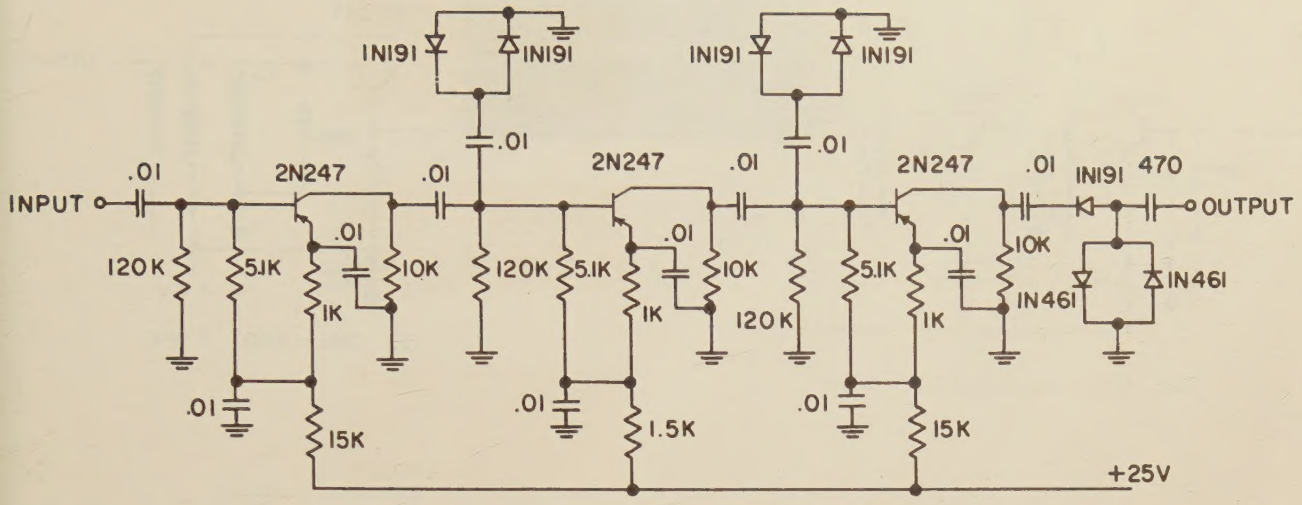


Fig. 12 - Video Amplifier Tested

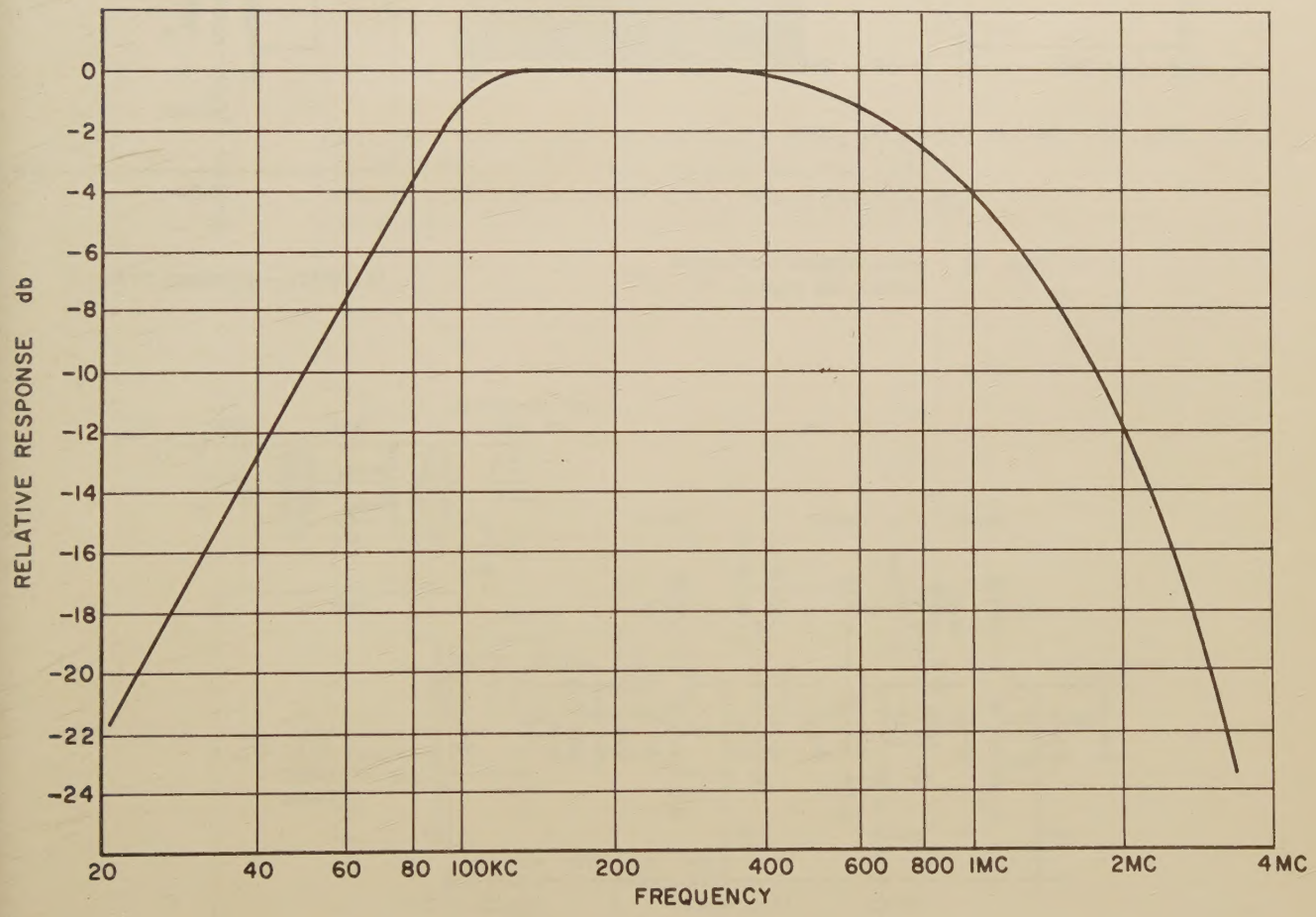


Fig. 13 - Response of the Video Amplifier

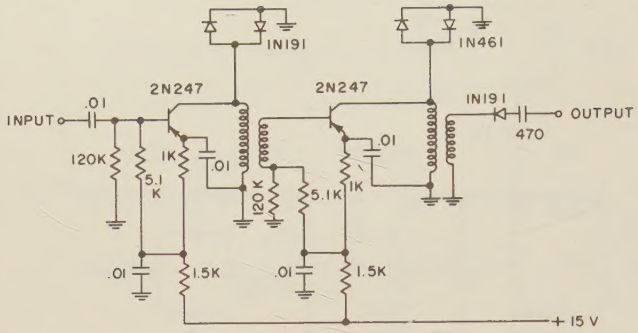


Fig. 14 - Pulse Amplifier Tested

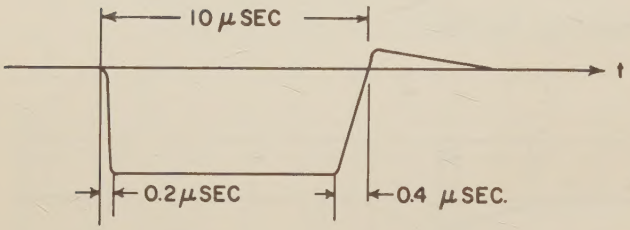
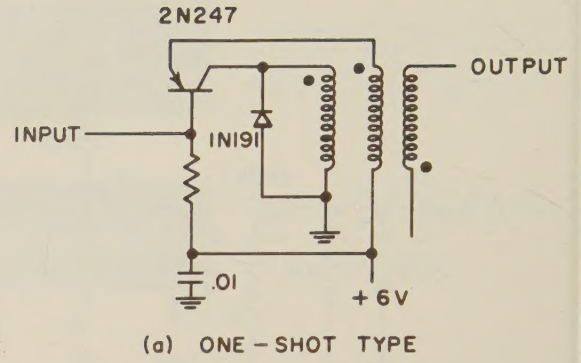
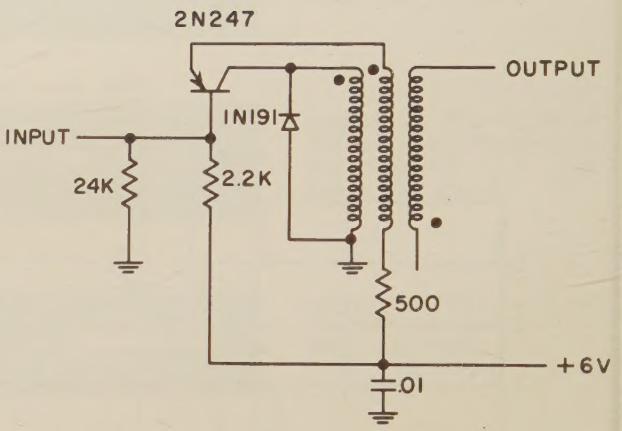


Fig. 16 - Gating Signal Waveform

Fig. 15 - Gate Signal Generator
(Photo at right)



(a) ONE-SHOT TYPE



(b) FREE-RUNNING TYPE

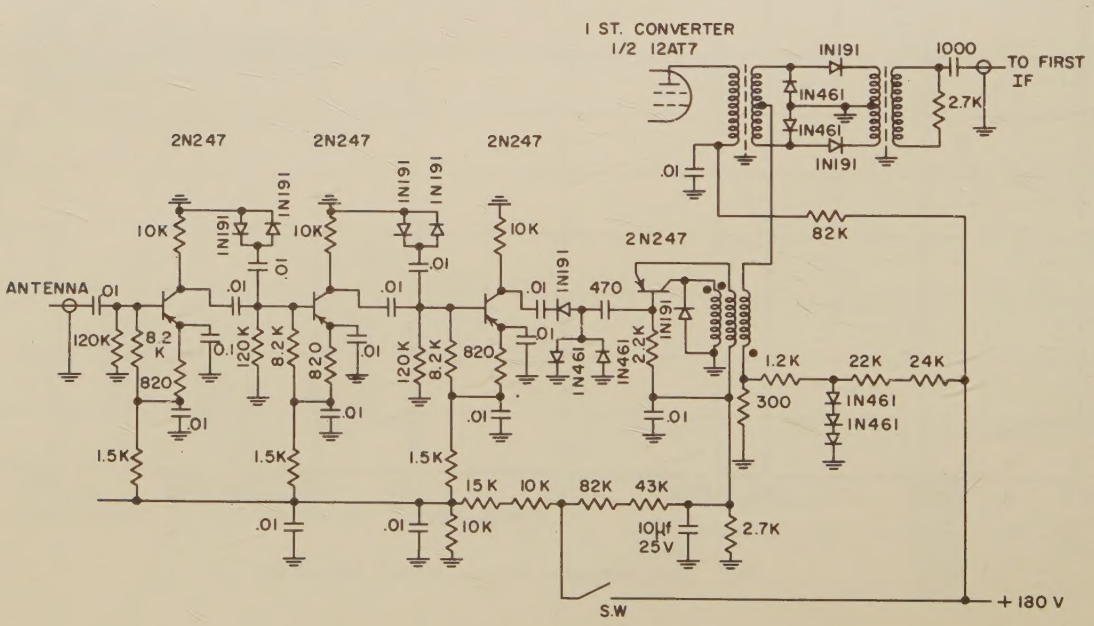
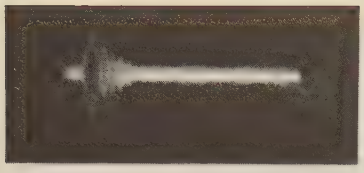
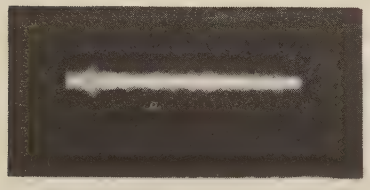


Fig. 17 - Impulse Noise Reduction Circuit



a

RECEIVER RESPONSE WITH NOISE REDUCTION CIRCUIT INACTIVE (AT THE PLATE OF THE FIRST LIMITER, MAXIMUM PEAK = 110V, 100 μSEC/ MAJOR DIVISION)



b

RECEIVER RESPONSE WITH NOISE REDUCTION CIRCUIT ACTIVE (AT THE PLATE OF THE FIRST LIMITER, MAXIMUM PEAK 20 V; 100 μSEC/ MAJOR DIVISION)

Fig. 18

A NEW CONCEPT IN THE RELIABILITY EVALUATION OF VEHICULAR RADIO

Frederick L. Hilton
Motorola, Inc.
Chicago, Illinois

Summary

This paper describes automatic equipment developed to conduct temperature tests of two-way vehicular radio. The automatic test unit reported below controls the temperature chamber, maintains input voltages at proper EIA specified levels, turns test specimens on and off, measures the operational performance parameters of transmitters and receivers at each of the selected temperature conditions and prints the data. Four units can be tested at a time and these may be any combination of standard frequency bands, carrier or continuous tone coded squelch, and wide band or split channel.

Introduction

It is well known that vehicular radio equipment is subjected to a variety of environmental extremes in its day-to-day operation. The Park Ranger in Death Valley expects year-around trouble-free operation from his two-way radio, and so does the Police Chief of Duluth, Minnesota. It is expected that vehicular radio will operate as well in New Orleans or St. Louis as in Phoenix, Arizona, as well on a veterinarians office table, as on the battery case of a forklift truck. The foregoing situations are not unique, but are part and parcel of the day-to-day operation of two-way vehicular radio. At Motorola, reliable performance at these extremes must be insured before a new design is accepted for production and for shipment to the customer. Consequently, each new design must be tested and re-tested under all conceivable combinations of these environments to insure its performance under such stresses. Such reliability testing is, of course, performed long before any units are shipped from the factory. First, the engineering units are tested, re-designed where the tests indicate changes are necessary, re-tested, and finally when performance measures up to requirements, the factory builds a pilot run of 50 or 100 units. Some of these are sent to the Design Engineering Dept., some to the Environmental Test Dept., and some to the Quality Control Group for evaluation of performance and adherence to standards of good engineering design and practice. All the environmental and field tests are performed again, and only when the product has demonstrated its capability to operate virtually anywhere at any time, is it approved for shipment to the customer. For example, on the Motrac product line, which was announced a year and a half ago, many thousands of manhours were expended in the environmental testing of engineering models and pilot run units in the 150mc band alone. These tests, which were conducted by the engineers in the Environmental Test Department, were in addition to the countless temperature, humidity, vibration, and field tests, which were run by the design engineers during the actual design process. Low band units were put through a similar reliability program.

Of all environments, temperature is one of the most critical. Unfortunately, temperature testing is the most time consuming of all environmental testing, due to the fact that the radio equipment must be conditioned several hours at each temperature prior to making the actual measurements. Since these conditioning periods may terminate at inconvenient hours at night or in the early morning, a typical temperature run on four units requires 3-1/2 days elapsed time. In addition, about 44 manhours are required to set up the test, make the measurements, and record the data. At the conclusion of the actual testing, additional manhours are required to organize the data and analyze it for reporting purposes. In view of the 3-1/2 day elapsed time requirement, and the extensive manhours, it was not feasible to test more than four units per week in a given temperature chamber, whereas it was desired to check many more engineering or pilot run units per week on new designs as well as a good sampling of current "standard production" units. It is not adequate to check the performance of pilot run units and then forget the product environmentally speaking, from then on. The performance of every product in the line must be checked at regular intervals to insure that nothing has slipped and that each item still measures up to performance requirements.

The Automatic Test Equipment

This situation was carefully reviewed to determine the best means for remedial action. Considered were such items as increasing the number of test chambers and test personnel, working around the clock with three shifts, and other such means to achieve our goal. As a result of this investigation, it was decided to develop automatic equipment which would conduct temperature tests, and in the summer of 1958 an engineering project was established for this purpose. The goal of this project was to design and install automatic, unattended equipment for performing and recording data from standard environmental temperature tests of two-way radio equipment. The following were the tests which were to be included:

Transmitter Tests

Power Output

Frequency
FM Noise
AM Noise
Modulation Sensitivity
Maximum Modulation Deviation
Tone Deviation
(for continuous tone squelch systems)
Battery Drain
Interstage Meter Readings

Receiver Tests

20DB Quieting Sensitivity
Squelch Sensitivity
(both threshold and tight squelch)
Sensitivity for Continuous Tone Squelch
Oscillator Frequencies
Audio Power Output
Distortion
Hum and Noise Level
Discriminator Stability
Battery Drain
Normal Interstage Meter Readings

In addition, it was desired to record internal temperatures at several locations in each of the radios to be tested. All tests were to be conducted in accordance with existing EIA Standards.

The unit to be described meets all the foregoing requirements and will automatically make complete temperature test runs on four separate radio units. The radios under test may be any combination of standard frequency bands and carrier or continuous tone coded squelch systems. Either the receiver or transmitter tests, or both, may be individually selected for each radio, but these are selected as a group, no individual breakdown of tests is provided. The automatic test unit controls the temperature chamber, maintains input voltage to the radios at the proper EIA specified levels, turns the test specimens off and on, measures the operational performance parameters at each of the selected temperature conditions, and prints the data. At the outset of the engineering project, it was decided that commonly accepted test methods would be used throughout. This complicated the design of the automatic test unit somewhat, but the complication was considered necessary from the standpoint of easy correlation of data with previously performed tests, and to keep the test results in a form familiar to those analyzing the data. Extensive use is made of servo systems for such purposes as adjustment of squelch and volume controls, signal generator tuning, output level adjustment, etc..

The completed automatic unit, nicknamed MASTER, the letters standing for Motorola Automatic Sequential Test Equipment - Radio, is shown in Figure 1. This equipment is housed in five racks. It consists of a "Programmer" chassis, a "Set and Test Selector" chassis, a "Receiver Test" chassis, and a "Transmitter Test" chassis. In addition, numerous items of standard test equipment i.e. signal generators, wattmeters, frequency counters, etc. are included. A control panel is provided where the operator programs the tests for each of the four units and also programs the type of temperature test to be run.

In operation the "Programmer" chassis functions to:

1. Turn the test specimens on or off at the proper time and in proper sequence.
2. Initiate the testing.
3. Control the temperature chamber.

The "Set and Test Selector" chassis, upon signal from the "Programmer":

1. Connects the proper radio to the measuring equipment.
2. Initiates the receiver test program.
3. Initiates the transmitter test program.

The "Receiver Test" panel conducts the receiver tests after receiving its cue from the "Set and Test Selector" and in a similar manner, the "Transmitter Test" panel conducts the transmitter testing. Standard test instruments are employed to make the measurements. For example, Measurements Model 210 signal generators, servo controlled, are used as a signal source. Data are measured with an Electro Instruments digital voltmeter which is coupled to a Hewlett-Packard Printer. Frequencies of the various oscillators are measured with a Berkeley frequency counter which is also connected to a printer, and temperatures are recorded on a Weston 24 Channel Temperature recorder. Figure 2 shows a close-up of the first two cabinets. The cabinet on the left contains from top to bottom, a distortion analyzer, the converter for the frequency counter, a panel upon which the frequency input information for each set to be tested may be programmed, the counter itself, the 24 channel temperature recorder, and some miscellaneous power supplies. The second cabinet contains the digital voltmeter, the main program panel, print control panel, failure alarm circuits, cycling circuits, and the relays and circuits for the main "Programmer" chassis and "Set and Test Selector" chassis. The third cabinet, shown in Figure 3, contains the receiver and transmitter test panels, meter reading circuits, and two printers, one for the digital voltmeter, the other for the frequency counter. At the bottom of this cabinet are the servo-controlled variacs which regulate the output from the battery chargers to maintain constant input voltages to the radio set cables in accordance with EIA standards. Cabinet four, Figure 3, contains four Bird Wattmeters, the 450MC modulation monitor, the control head mounting fixtures for two of the four control heads, the 30MC and 150MC signal generators, and battery chargers. Figure 4 shows

a closeup of the servo-controls and mounting fixtures for the control heads. Cabinet five, Figure 5, contains coaxial switching relays, the 150MC and 40MC modulation monitors, the 450MC signal generator, the mounting fixture for the other two control heads, a panel on which the input voltage information is programmed and battery chargers. The batteries, eight in all, are mounted in racks outside these cabinets, between the MASTER unit and the temperature chamber. Figure 6 shows a typical unit prepared for installation in the test chamber. In addition to the normal power and control cables you will note the presence of metering cables, pick-ups for both receiver oscillators, a cable which injects a 455KC signal and the 6 thermocouples. The unit in this picture is shown with covers removed to demonstrate the placement of cables, but in a normal test the covers are in place.

Let us now follow through a portion of a typical test program. Four Motrac two-way radios; two in the 150MC band, and two in the low band are being tested, one of each of these being units with continuous tone control squelch. The first unit is installed as set #1 and reference to Figure 7, which shows a close-up of the main program panel will show that the frequency band switches to indicate 150MC have been set, the switch indicating that the unit is a split channel radio has been set, and both receiver and transmitter tests have been programmed. Similarily in going down the panel the other three units, sets no. 2, 3, and 4 have been programmed. A temperature cycle of room temperature, -30°C, +60°C, and back to room temperature has been selected. In addition to the program on this panel, the frequency selector switches in cabinet 1 have been properly set so that the frequency counter will be tuned properly, and the input voltage switches in cabinet 5 have been set so that EIA specified input voltages will be constantly maintained during receiver and transmitter testing. Having programmed all the necessary information, and installed the proper cam on the temperature controller of the test chamber, the test program is started by pushing the "Sets 1 to 4" start button. When this button is depressed, set 1 is turned on immediately, and is permitted to warm up in the standby condition for a 15 minute period, after which set 2 is turned on and the test cycle is initiated on set 1. Upon initiation of the test cycle, the squelch control is turned all the way down to a wide open squelch position and the volume control is turned up to standard noise output level. Following this, the receiver metering stepping relay connects the digital voltmeter in turn to each of the eight receiver metering positions. At each position, the digital voltmeter is permitted 7 seconds to stabilize, and then 3 seconds to print the data. The tape is coded with the set number and also information as to which test the particular line of data is related.

After completion of the meter readings, the squelch control is adjusted to threshold by its servo motor and the signal generator is tuned to the receiver frequency. Figure 8 shows the signal generator with its servo drives and clutches. The first step which occurs is that the generator output level is adjusted to approximately 100 microvolts. This level is determined by a cam on the attenuator shaft. A bias voltage is then fed to the servo amplifier which drives the signal generator frequency up scale. When the receiver is "quieted" a relay closes, which switches the 2 speed clutch to "slow" and at the same time the input to the tuning servo amplifier is switched to the discriminator of the specimen receiver. Thus, the generator is zeroed in on frequency. A time delay relay is employed which, after giving the generator tuning motor time to have reached the high end of the band, initiates the next part of the program. The attenuator is driven down scale to zero output and then up scale until the squelch just opens, the noise output being the criteria for this measurement. The digital voltmeter is connected across the output of a logarithmic precision potentiometer which is geared to the attenuator shaft. This potentiometer is supplied with a standard input voltage and its output is directly proportional to the signal generator output so arranged that 100 millivolts is equal to 1 micro-volt output from the generator. Thus, the digital voltmeter prints a number which is equivalent to the generator output in microvolts.

Next, the squelch control is adjusted for maximum (tight squelch) and the signal level is raised again until the squelch opens, and this level is printed. Thus, the tight squelch sensitivity has been measured.

If the receiver employs a continuous tone controlled squelch system, the generator is modulated with the appropriate tone frequency at minimum specified deviation and again the signal is adjusted for squelch opening. If the specimen radio is a carrier squelch unit, this step is omitted.

The next test is for 20 db quieting sensitivity but before this measurement is made, the output from the generator is raised to 100 microvolts and the input to the generator tuning amplifier is reconnected to the discriminator so that the generator checks its tuning just prior to the 20 db quieting measurement. For this measurement the squelch control is once again turned to minimum squelch and the signal level is increased until 20 db quieting has occurred. The digital voltmeter is connected to the precision potentiometer and the output is measured as before. Subsequent to the sensitivity testing, the output level of the signal generator is raised to 1000 micro-volts for audio measurements. The first of these is measurement of audio power output and for this measurement the signal generator is modulated with standard test modulation and the digital voltmeter is connected across a 3 ohm load. Next, the audio distortion is measured at this level by

switching the digital voltmeter to the "oscilloscope" terminals of a distortion analyzer. Following this measurement, the modulation is removed and the voltmeter measures the hum and noise level. At this point, the output level of the signal generator is turned down to zero and the generator is tuned back to the low end of the band. During the time that this out-of-band tuning is in process, a standard 455KC signal is injected into the back IF and the discriminator reading is recorded as an indication of discriminator stability. Finally, the standby drain to the receiver is measured by connecting the digital voltmeter across a standard shunt in the "A" power lead.

In addition to the measurements which have just been described, the frequency of the oscillators in the receiver is being measured and printed during this time. The frequency counter is first connected to a probe near the second oscillator, then to the first oscillator.

Upon completion of the receiver tests, the transmitter test cycle is initiated. The transmitter test program will not be described in detail, but it follows the same format as used in the receiver test program. Meter readings are taken and other measurements are made by connecting the digital voltmeter to the appropriate place at the appropriate time. The transmitter testing is concluded with measurement of the transmit drain.

This completes the measurement data on set 1 except that if this set had been a 2 frequency unit, it would now be switched to channel 2 and the measurements would be repeated. The main programmer takes cognizance of the measurement time required and waits to turn on the second radio until there is only 15 minutes time remaining before it is to be tested. The test program is conducted on sets 1, 2, 3, and 4 in order, and upon completion of the tests on the last set, the "MASTER" unit goes into a "stand-by" condition and the temperature chamber is moved to the new ambient temperature. Upon reaching the new ambient, the equipment is allowed to stabilize for 5 hours as required by EIA Standards prior to the next measurements being taken. The test chamber requires slightly less than 2 hours to move from one chamber temperature to another, and this combined with the 5 hour soak time gives a total of 7 hours during which the "MASTER" unit is inoperative. Referring once again to the main program panel, Figure 7, we note that provision has been made for testing set 5. This input channel is available for use during the long standby periods. It is programmed exactly the same as sets 1 through 4 except that it takes 1 set of data (or 2 sets in the case of a 2 frequency unit) and shuts down. These tests which are initiated with the "Set 5" start button permit the complete testing of a radio which has just come from the humidity chamber or which has just come off the vibration table. When the 7 hour elapsed time has been completed, the MASTER unit automatically begins testing the 4 original units in the temperature chamber, even though the test of unit 5 might still be in progress. In order to avoid this situation, a clock is provided on the control panel. This unit runs "as a clock" during all but the standby periods. At the beginning of the standby period, the clock motor is stopped, its hour dial is advanced 7 hours, and a lighted sign comes on stating "the next test cycle will start at" and the clock then shows the time when the next test cycle will begin. When the 7 hour delay time has elapsed, and the test program starts, the clock reverts to being a clock.

The test programs at -30°C are carried out exactly as previously described, but at +60°C the EIA Standards prescribe certain cycling of the transmitter prior to measurement of either receiver or transmitter. The "MASTER" unit at +60°C turns on all the specimen radios and cycles them 1 minute on transmit, 4 minutes on standby during the prescribed 5 hour period, and at the end of this time, takes measurements on each of the units. In this case, the receiver measurements are made during the 4 minute standby periods. Transmitter measurements are made by keeping one transmitter on during the 4 minute period when the other 3 units are on standby so that the other transmitters do not interfere with the measurements. At the conclusion of these high temperature tests, the unit would normally move to the next ambient, soak the equipment, and continue the testing. However, if the "Temperature Rise" tests have been programmed, the transmitters are cycled for an additional 3 hours at 1 minute on, and 4 minutes standby and then during a final 9th hour, the transmitters are operated on a 5 minute on, 15 minute standby basis. At the end of this 9th hour of testing, transmitter power output, frequency, and meter readings are measured. Also, during these last four hours, the 24 channel temperature recorder is measuring temperatures at 6 locations in each of the 4 units.

Provisions have been made to recognize the failure of a piece of equipment which has been on test. The failures which are recognized are generally of the catastrophic type. When such a failure is recognized, a failure light on the program panel indicates that a particular receiver or transmitter has failed, and from then on, the test program for that particular unit is skipped just as though no unit had been programmed in that spot in the first place.

Upon completion of all testing, the MASTER unit shuts down and waits for the operator to return to analyze the data. Should we desire to run some special test on the specimen radio at a given temperature extreme, provision is made to interrupt the regular test program and hold all units at this extreme until the desired tests have

been made. Upon completion of these tests, turning off the hold switch permits the test program to continue.

The data obtained from these tests are printed on 2 tapes which are shown in Figure 9. The frequency tape is shown on the left, and in this case, the last digit to the right is employed to identify the unit under test. For example, the number (1) in this column signifies set no. 1 receiver, (2) signifies set no. 1 transmitter, (3) set no. 2 receiver, and so forth. This tape can easily be read since the only information thereon is a 1 digit "set identifier" and the oscillator frequency in cycles. F1-F2 information can be obtained directly from the frequency printed. The digital voltmeter tape shown on the right in the figure is much more difficult to read since it includes a variety of receiver and transmitter tests where the data printed may be in microamperes, microvolts, volts, or amperes. To assist the test engineer in reducing the data from this tape, we have developed a tape reader. This device, which is shown in Figure 10, is used to read the digital voltmeter tape. Referring to the figure, note that the individual receiver tests are listed at the left, and that each test has an identifying number. Through the various windows in the reader, portions of the tape may be seen. Looking at the left hand column of numbers on the tape, we see the set and frequency identifier. In this case, a number 1 represents set 1, frequency 1, number 2 represents set 1, frequency 2, and so forth. The next 2 columns indicate the particular test which was taken. Note that numbers in this column must agree with the numbers printed on the reader near the test identification at the left. Due to the manner in which the data comes from the printer, the order of tests is read from bottom to top. The next column is to indicate the polarity of DC signals and also, the sine wave symbol in this column indicates that an AC parameter was measured. The actual data are printed in the next several columns with the vertical lines indicating the location of the decimal point. If we look near the bottom of the chart, we will see that receiver position 2 is reading 16.0. This is the same reading which would be obtained if we had plugged a standard Motorola test meter into the receiver and read its meter reading. All meter positions, 1 through 8, are recorded. Test number 09 for threshold squelch sensitivity is provided with 2 possible decimal lines. If a 4 appears in the small window at the right, decimal line 4 is used, and similarily, if a 3 appears in the small window at the right, decimal line 3 is used. This particular reading is presented in microvolts and thus we see that the threshold squelch sensitivity for the particular set being tested is 0.13 microvolts. The maximum squelch sensitivity and 20 DBQ measurements are printed in the same manner. The audio power output distortion and hum measurements are presented as AC voltages. Audio output is seen to be 3.92 volts, which can be squared and divided by the output load impedance to indicate power output in excess of 5 watts. The distortion measurement is so arranged that a 10-1 multiplier is employed, and consequently, a position 14 reading identical to the position 13 reading would indicate 10% distortion. The remaining data are shown recorded on up the chart.

The tape reader is turned over to read transmitter data and this portion of the reader is shown in Figure 11. The method employed is exactly the same as was employed on the receiver test presentation, except that you will notice that the transmitter tests have been coded, starting with no. 30.

The equipment, which has been described, was designed to automatically test standard two-way mobile radio equipment in the three commonly used frequency bands. The present unit does have limitations, for example, it could not be used to measure the performance of a 900 MC transistorized receiver used in our traffic control equipment. Such measurements must still be made manually.

We often use the phrase "in accordance with standards of good engineering practice". The periodic review of the environmental performance of a product most certainly falls within the framework of "good engineering practice". The equipment we have discussed here-in provides the means to constantly evaluate the temperature, humidity, and vibration performance of our product on a weekly sampling basis. Thus we are able to insure that the performance of the 100th, 1000th, or 10,000th unit is equivalent to the performance of the engineering design model.

Fig. 1 - General View of Complete "MASTER" Unit

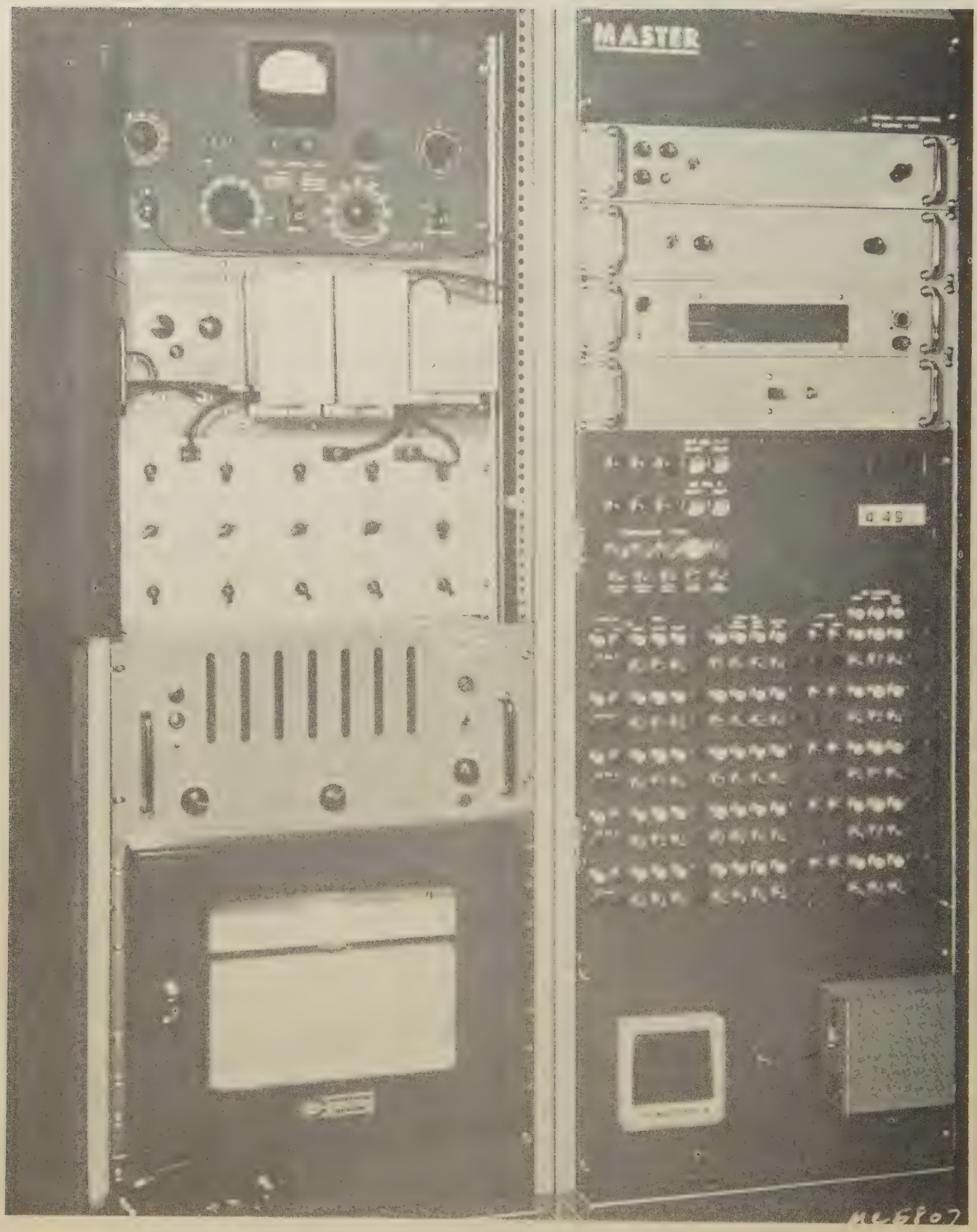


Fig. 2 - Close-up View of Cabinets 1 and 2

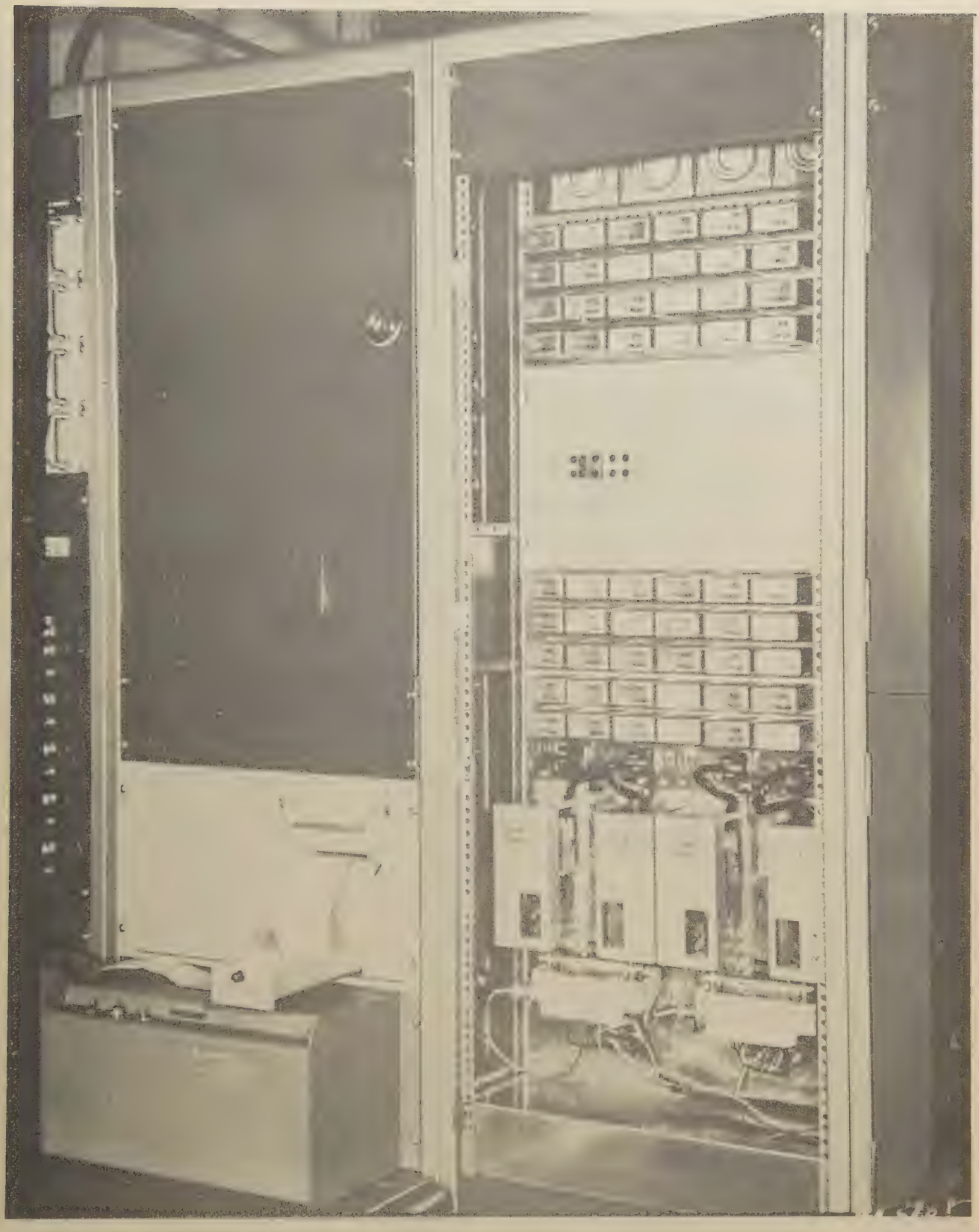


Fig. 3 - Close-up View of Cabinets 3 and 4

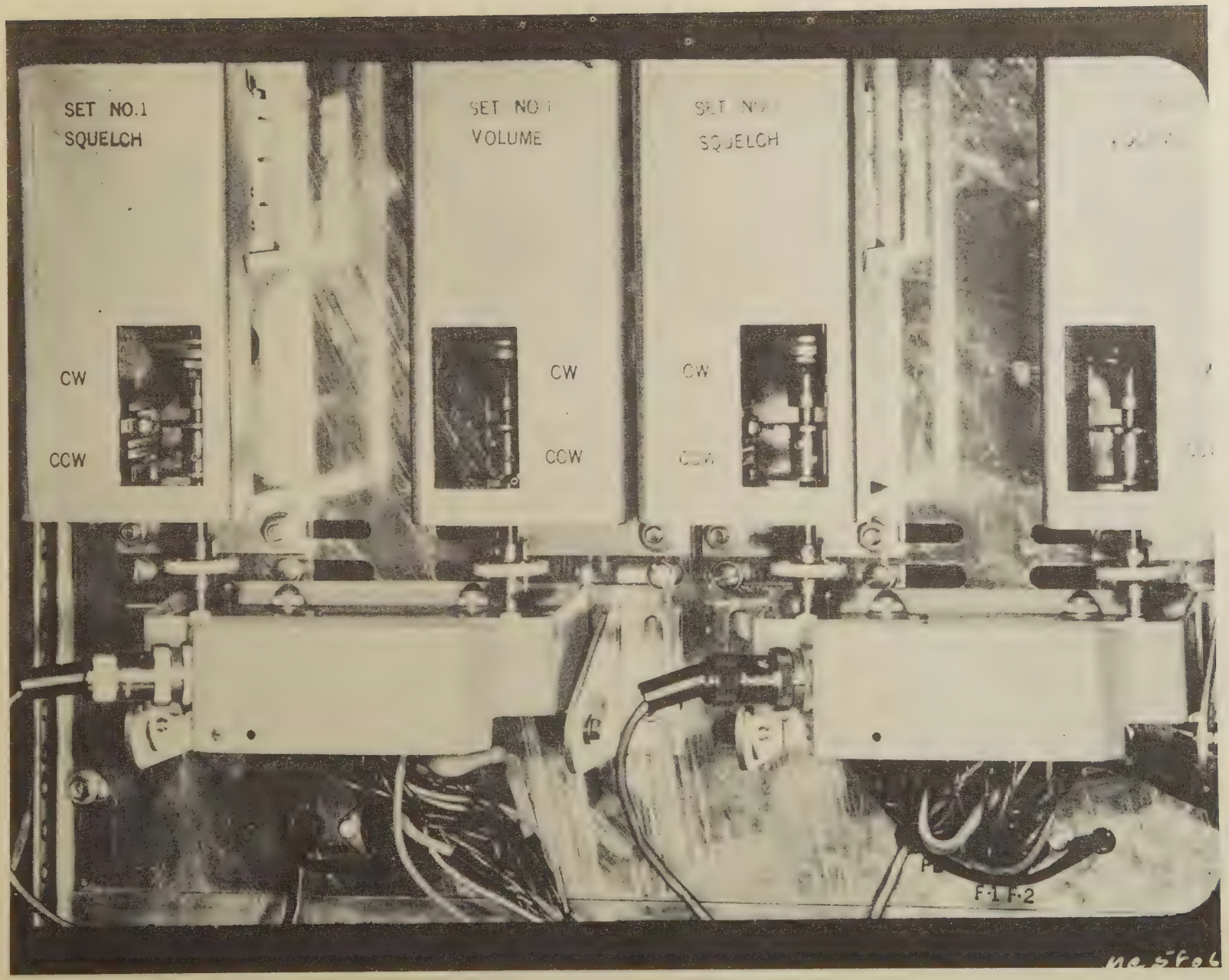


Fig. 4 - Close-up of the Control Head Servo's

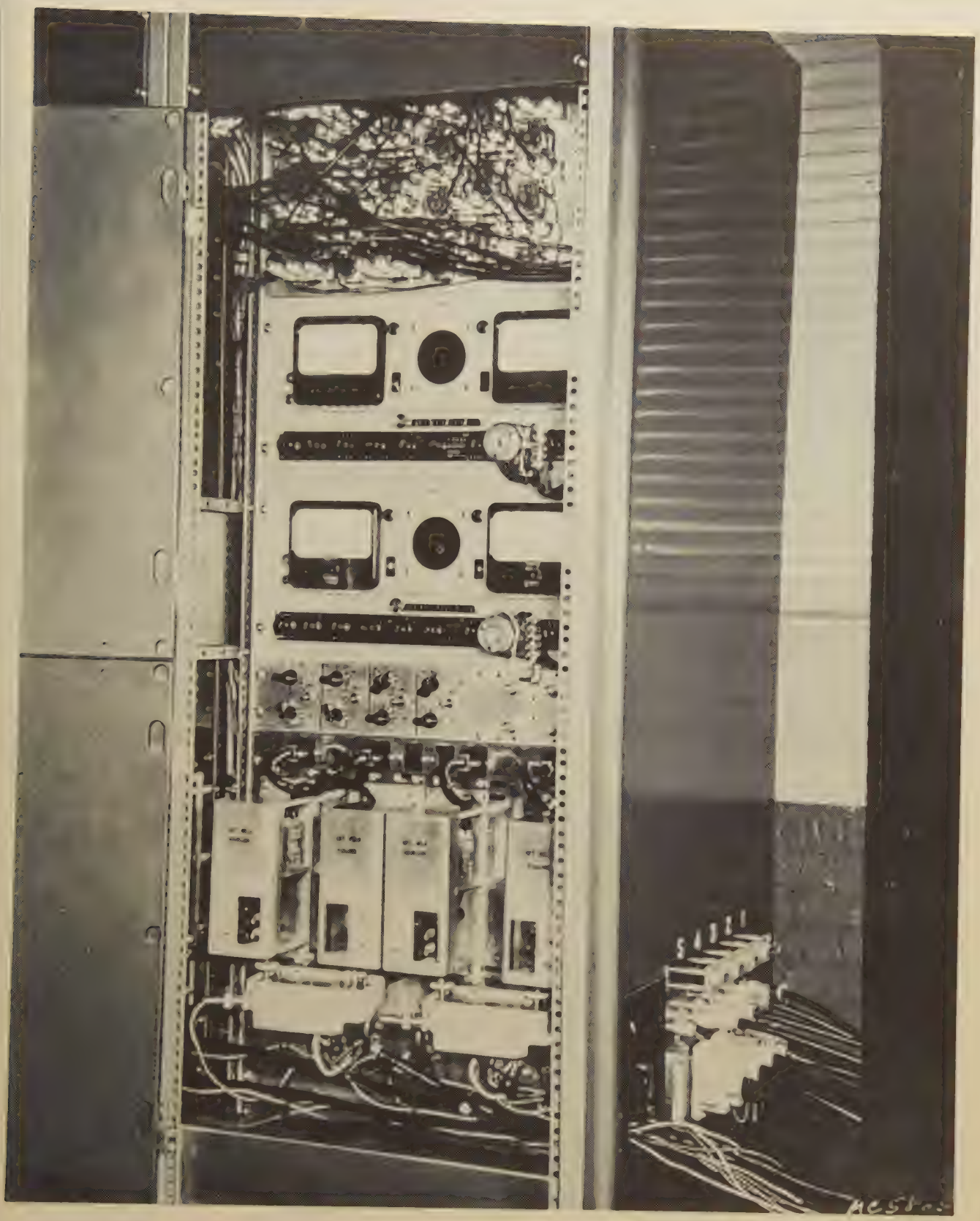


Fig. 5 - Close-up of Cabinet 5

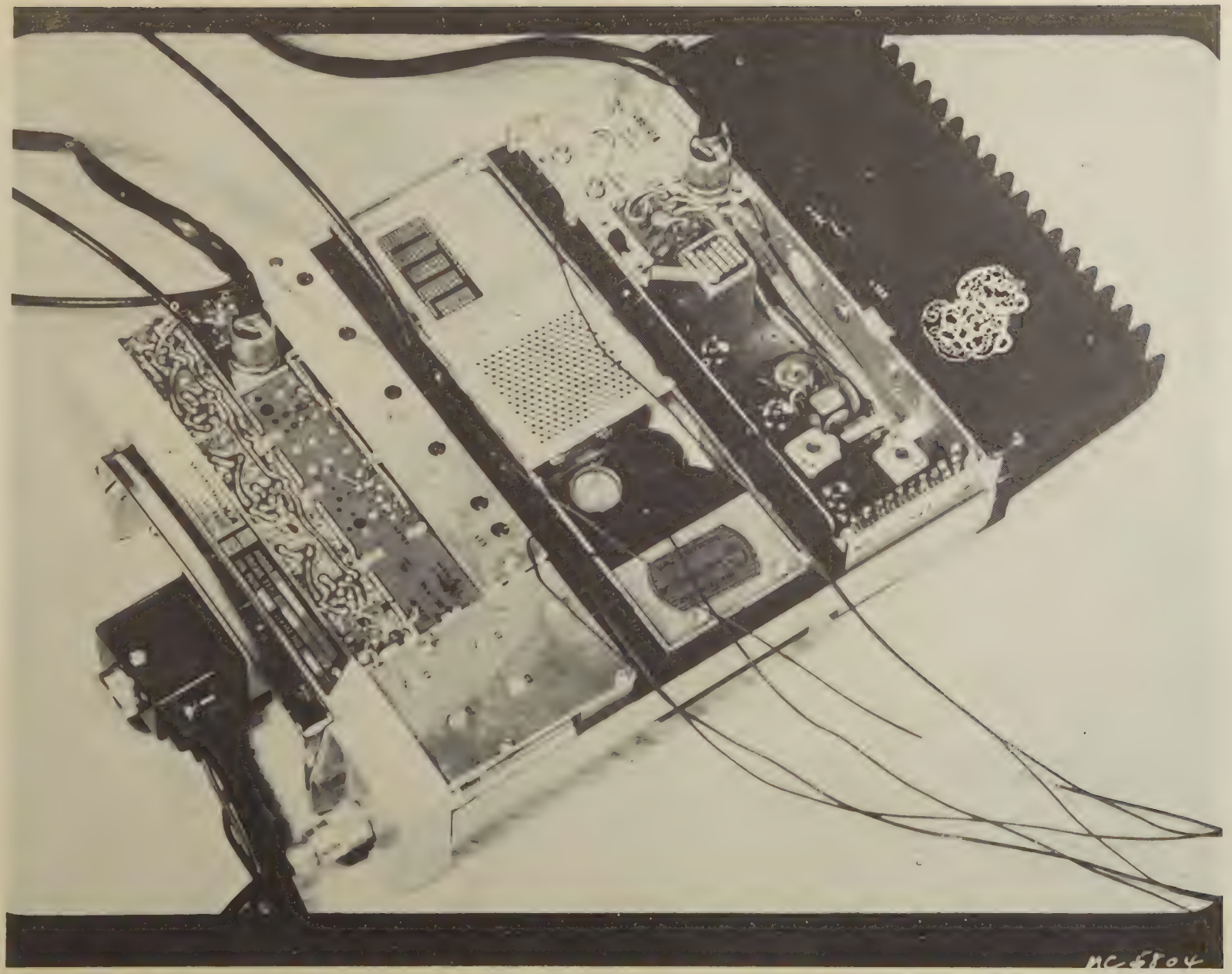


Fig. 6 - Close-up of Test Specimen with Meter Cables, Frequency Pickups, and Thermocouples

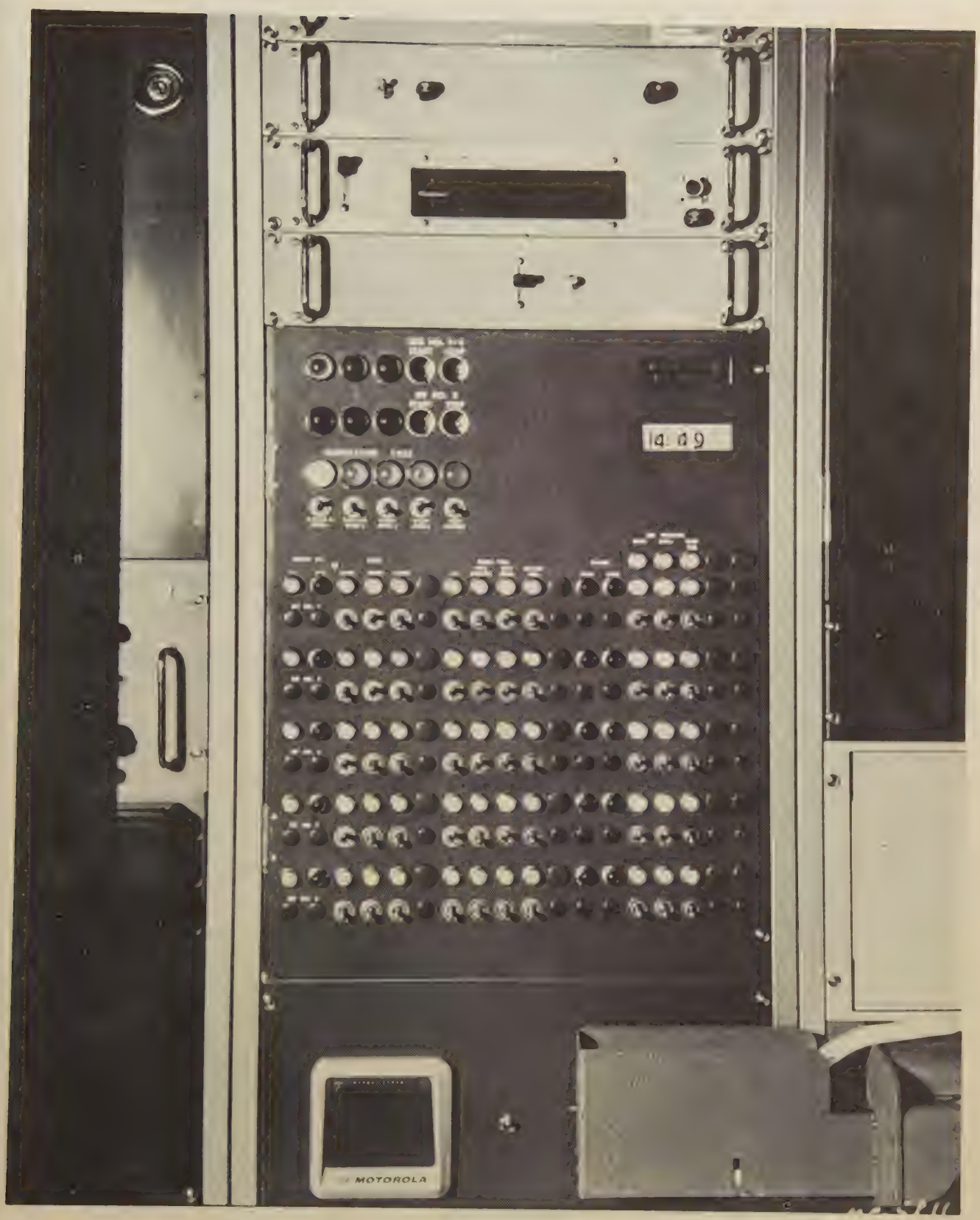


Fig. 7 - Close-up of Main Program Panel

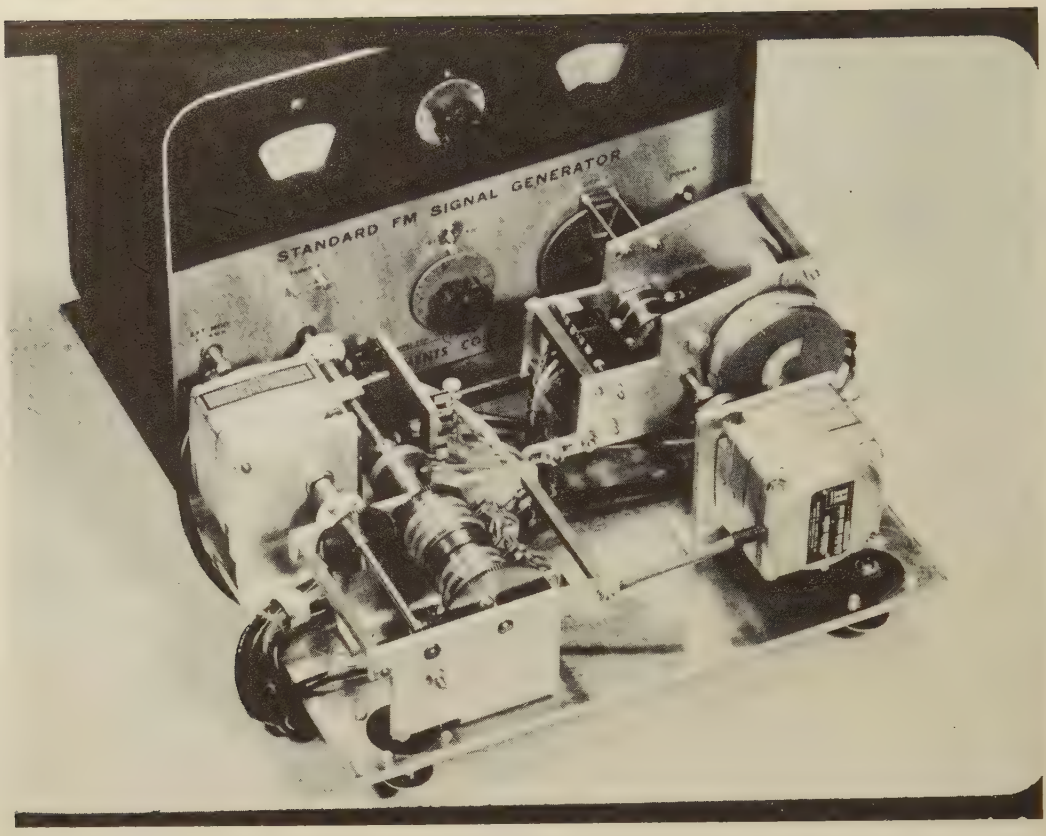


Fig. 8 - Signal Generator Equipped with Servo Drives

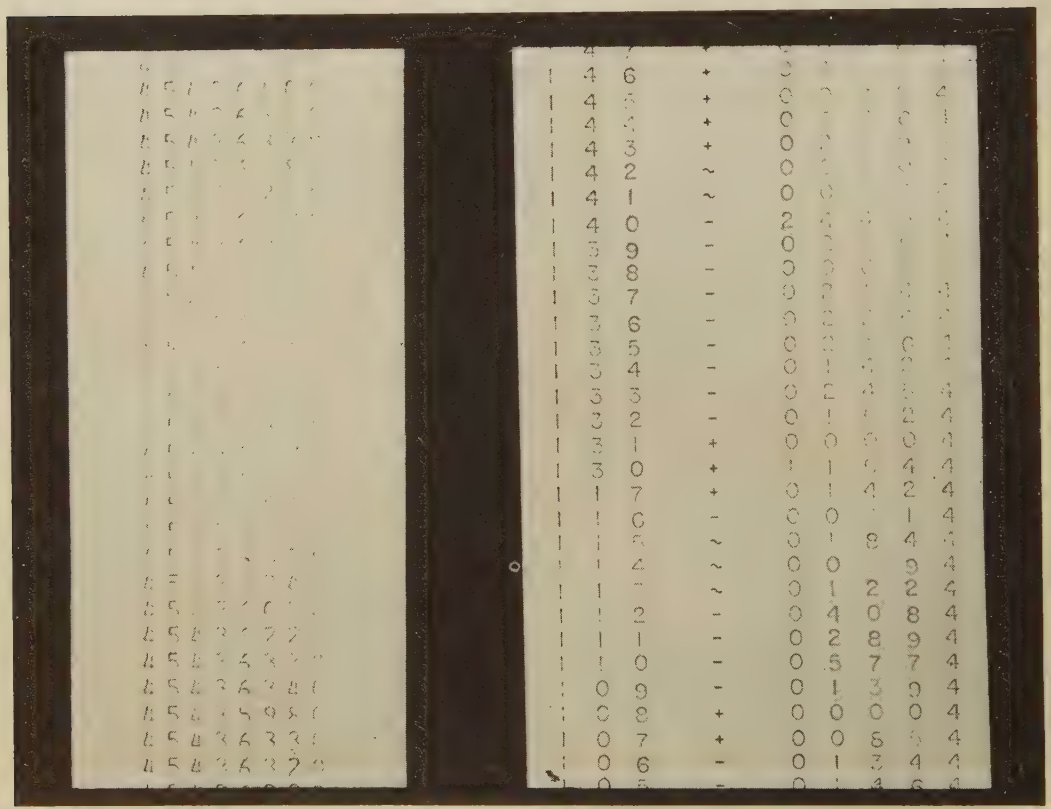


Fig. 9 - "MASTER" Unit Data Tapes

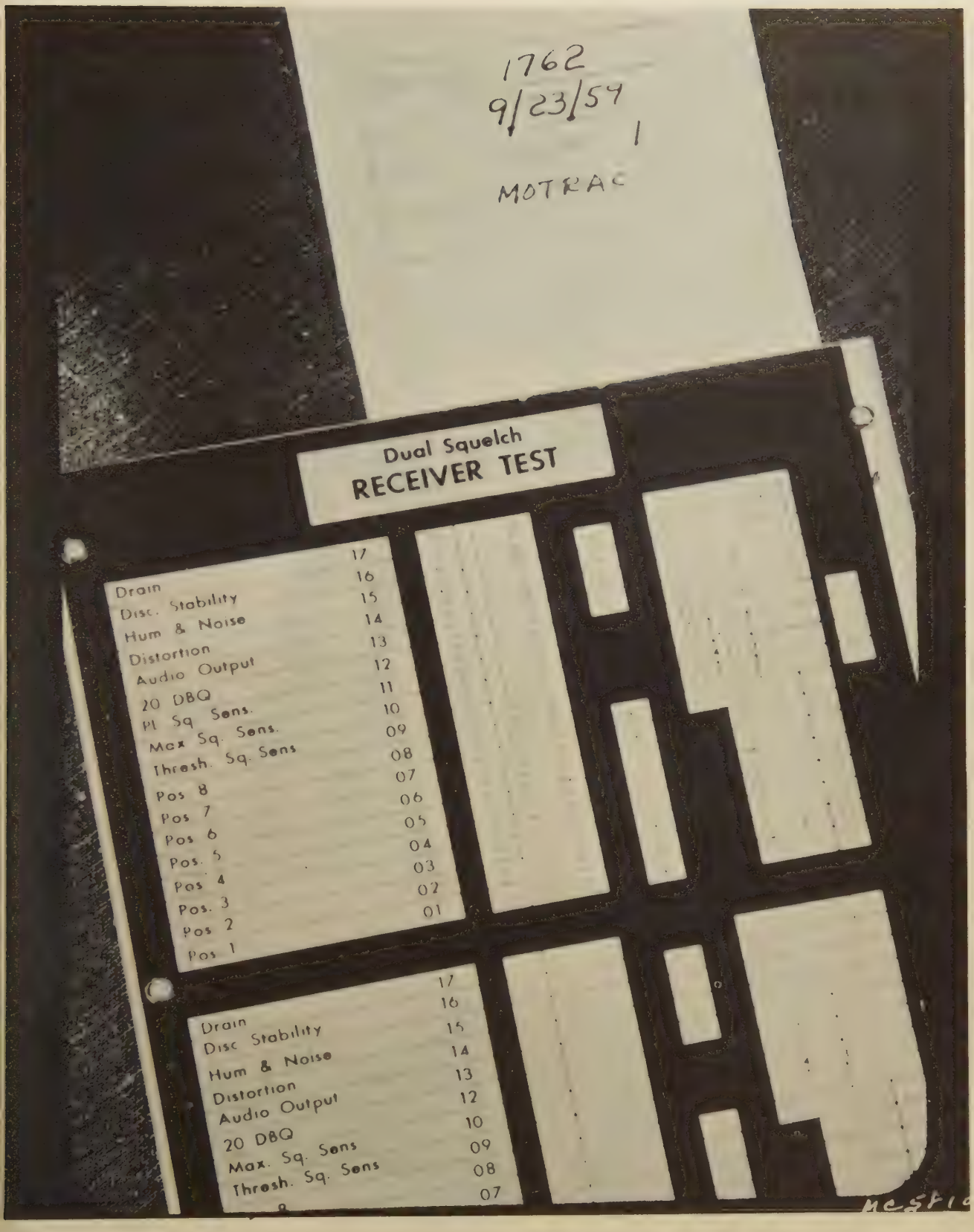


Fig. 10 - Receiver Data Tape Reader

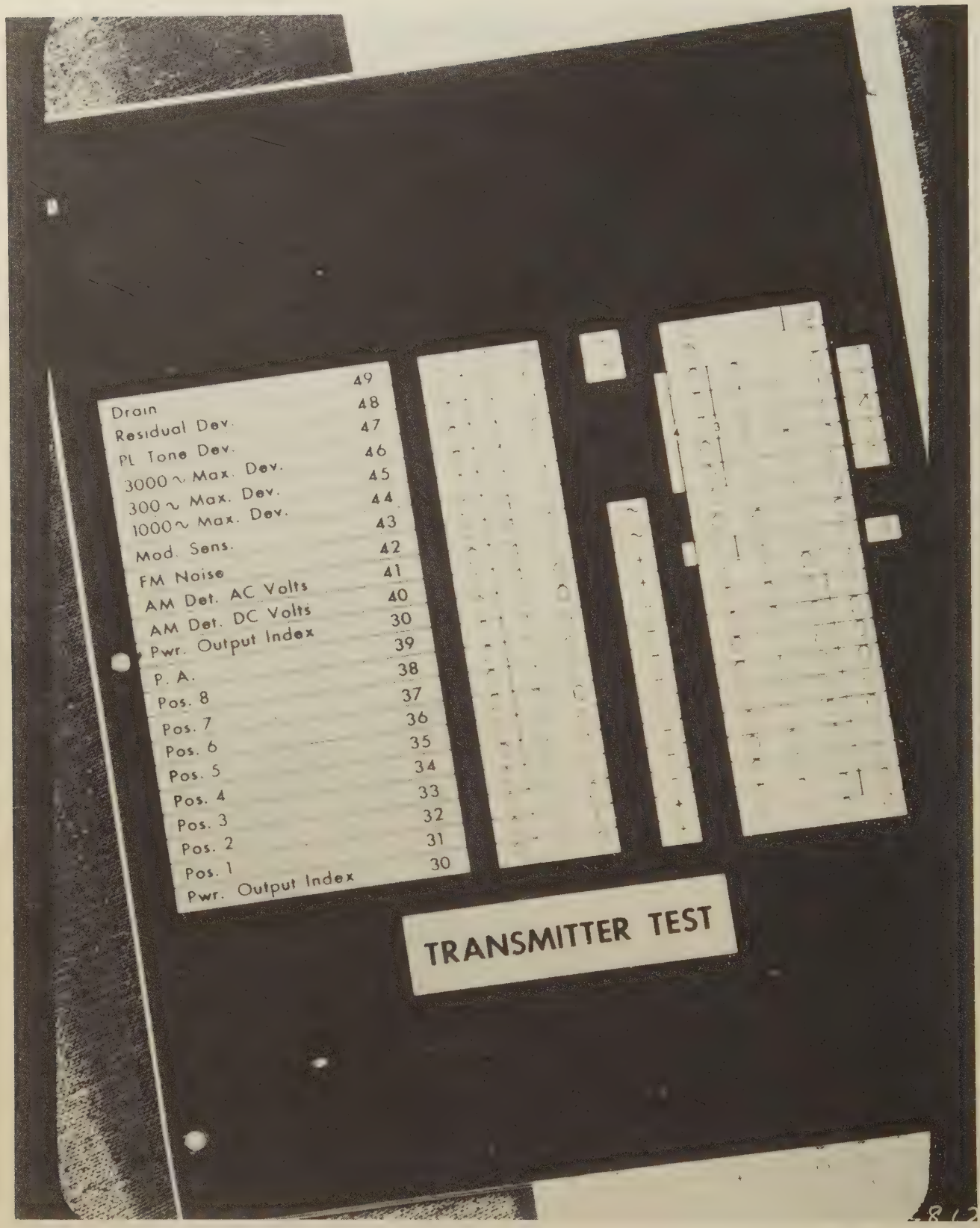


Fig. 11 - Transmitter Data Tape Reader

THE NEW TREND IN MINIFIED COMMUNICATIONS EQUIPMENT*

J. W. Knoll†, Senior Member, IRE

Summary - This paper presents a brief description of the Micro-Module building-block and the current program status in terms of both up-to-date hardware accomplishments and future consideration, particularly in the area of miniaturized communications equipment. The advantages offered by the revolutionary Micro-Module are reviewed and illustrated with reference to size reduction and packaging design potential in the following:

(1) AN/PRC-36 Combat Helmet Radio Set
(2) AN/URC-11 Personal Rescue Receiver-Transmitter
(3) AN/PRC-25 Portable Radio Set
(4) HF Single Sideband Receiver
(5) Non-Military and Commercial Communication Equipment

INTRODUCTION

The present day packaging design trends, focused by recent space accomplishments, emphasize the serious need for high density equipment possessing a high order of reliability and "common sense" design. There can be no doubt as to the importance of further design effort toward miniaturization and weight reduction of payloads for missiles and space vehicles, as well as for ground vehicular type electronic equipment. The field of RF communications circuitry provides, perhaps, the greatest opportunity for immediate success for minified equipment. Present day portable RF communication equipments are still too large and bulky —— these must be dimensionally reduced and streamlined to fit the size requirements soon to be in demand.

The RCA Micro-Module Program, under Signal Corps Contract, is providing the military with a new and broad production potential for fully reliable micro-miniature electronic equipment, in which the Micro-Module can serve as a standard building block. In February of this year, the first live demonstration of a number of ground communications equipments incorporating the Micro-Module was presented to the Department of Defense. Here, the validity of the Micro-Module was clearly established. Emphasis was placed on its applicability not only to the areas set forth under the present military program but to the whole future of solid state techniques, including various commercial applications.

* Original manuscript received by the IRE, December 15, 1959. This work was supported in part by the U.S. Army (Signal Corps) and in part by Radio Corporation of America.
† RCA, Defense Electronic Products, Surface Communications Division, Micro-Module Engineering, Camden (2), New Jersey.

MICRO-MODULE BUILDING BLOCK

Now let us look at this Micro-Module, Figure 1.

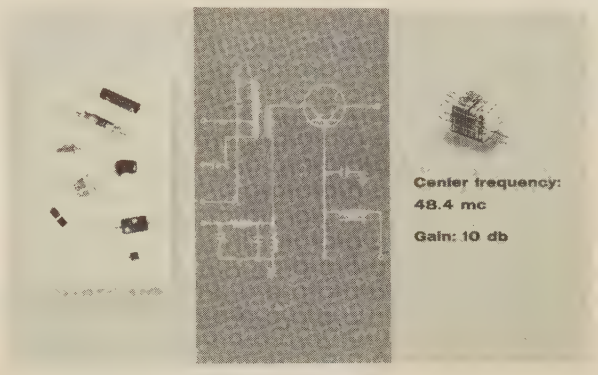


Fig. 1 - RF Module

Shown here is a schematic of an RF amplifier. In a conventional piece of equipment, this circuit would be constructed physically by assembling the handful of electronic components shown at the left in accordance with the circuit diagram. Note that these conventional components have a variety of sizes and shapes; they differ widely in space requirements and mounting problems. The Micro-Module approach to this same circuit is shown at the right where the complete circuit function is contained in a stacked group of wafers called micro-elements, each of which is electrically equivalent to a conventional component. All interconnections within the circuit are made through twelve "riser-wires" which are soldered to the edges of the wafers. The same wires also serve as means for making, input, output, and power connections to the circuit as a whole. The entire module is encapsulated in epoxy to form a compact, rugged assembly.

The key technical feature of the Micro-Module Building Block is "disciplined geometry", Figure 2. All are in the form of square wafers approximately 3/10's of an inch on each side, which can be stacked and interconnected in a wide variety of ways dictated by circuit design. The assembled Micro-Module is then of uniform cross-section, facilitating extremely compact module assemblies.

With the Micro-Module we have produced reliable electronic equipment reduced in size by a factor of 10 as compared to equipment using conventional components. The Micro-Module structure

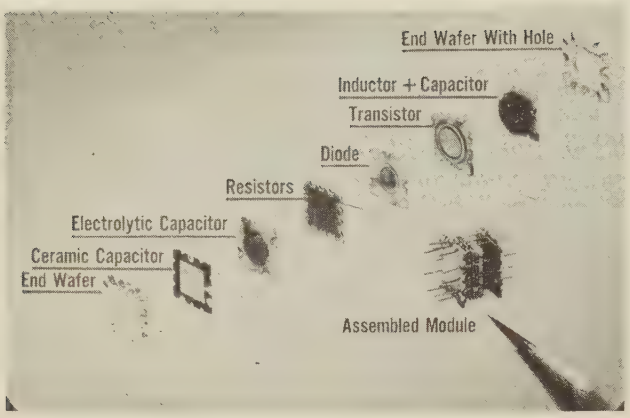


Fig. 2 - Typical Micro-Module (Exploded View)

provides a logical "frame-work" to include many of the highly advanced solid-state circuit techniques now being explored in many leading research laboratories.

RCA is exploring the possibility of utilizing tunnel diodes, unipolar transistors, and complete integrated circuits into single-wafer form such as a full adder, Figure 3. Recent laboratory successes in this area show read promise for achieving ultimate equipment size reductions in excess of 100 to 1.

In addition, the Micro-Module structure provides the means of incorporating such concepts as Texas Instruments Integrated Circuits and Westinghouse's Molecular Electronics.

PRESENT AND FUTURE DEVELOPMENTS

Now let us examine where the Micro-Module Program stands today in terms of both hardware accomplishments and some proposed design considerations. This presentation will cover various developments, packaging schemes and related size reduction efforts of the following:

AN/PRC-36 Combat Helmet Radio Set
AN/URC-11 Personal Rescue Receiver-Transmitter
AN/PRC-25 Portable Radio Set
HF Single Sideband Receiver
Non-Military and Commercial Communication Equipment

The transistorized AN/PRC-36 Combat Helmet Radio Set, represents a highly miniaturized equipment using conventional components. Currently being tested by the military, this FM set is designed for operation at any one of 12 discrete frequencies in the range of 37 to 51 megacycles. The receiver portion of this radio set has been packaged in Micro-Modular form and successfully demonstrated.

In this case, the Micro-Modular Receiver, Figure 4, was designed to clip on the side of a combat helmet. The unit contains a complete single frequency super-heterodyne FM receiver with batteries and accessories. It weighs 12 ounces

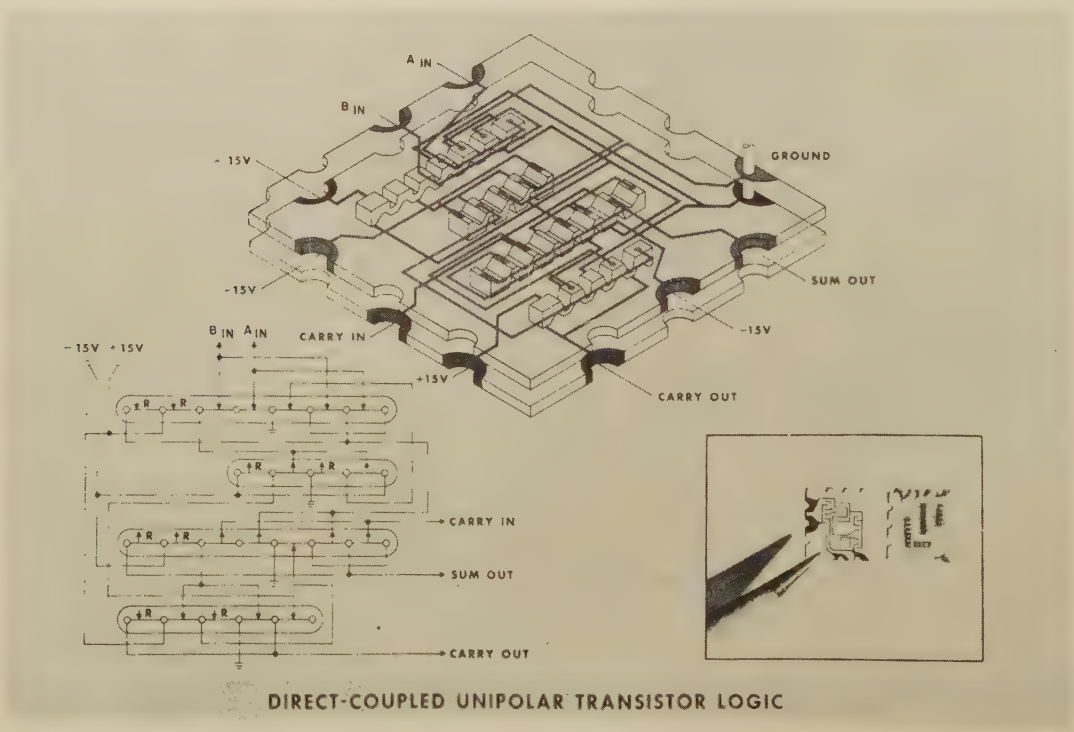


Fig. 3 - Micro-Module Full Adder

and occupies a volume of 3 cubic inches. Included in this unit are 11 Micro-Modules performing the following functions:

R-F Amplifier (1 Module)
I-F Amplifiers (5 Modules)
Audio Amplifier (1 Module)
Mixer (1 Module)
Crystal Controlled Oscillator (1 Module)
Limiter (1 Module)
Discriminator (1 Module)

The complete Helmet with the Micro-Module Receiver weighs 2¼ pounds, which is a "far-cry" from the 40-pound radio carried by the soldier on his back during World War II and through part of the Korean campaign.

It cannot be too greatly emphasized that Micro-Modular communication equipment such as this offers the size and weight reduction which can improve the over-all mobility and effectiveness of our ground troops.

A sophisticated version of the Helmet Radio Receiver, Figure 5, is currently in initial production at RCA. In this model a 5 to 1 size reduction and a packaging density of 250,000 parts per cubic foot are realized. Ten modules including the IF Amplifiers (four now considered sufficient) are mounted on end in two rows on the printed circuit board.

A simple and effective electrical layout, including shielding consideration, and a minimum number of parts result in a rigid compact structure. These factors all contribute to the high reliability.

As a second generation development effort, the Receiver-Transmitter portion of the Helmet Radio Set has been "phased" into the Micro-Module Program. Figure 6 shows a volumetric comparison of the conventional and proposed Micro-Modular equivalent of the AN/PRC-36 Transceiver. The conventional occupies 8.2 cubic inches, while the Micro-Module Transceiver is 2.8 cubic inches. The

Fig. 4.- Micro-Module Helmet Receiver

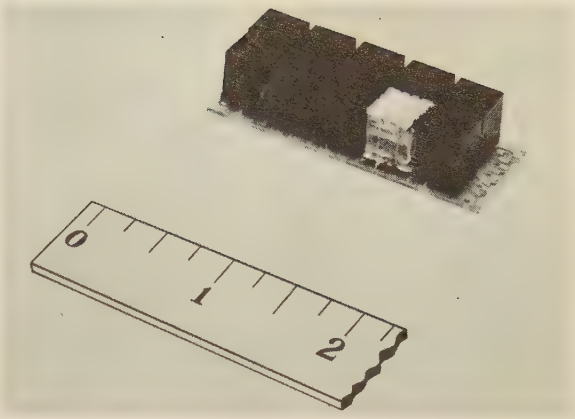


Fig. 5 - Micro-Module AN/PRC-36 Receiver (Production Version)

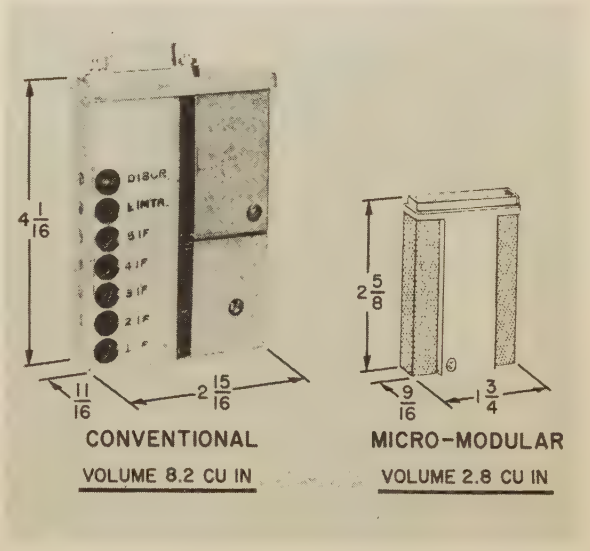


Fig. 6 - AN/PRC-36 Transmitter-Receiver.

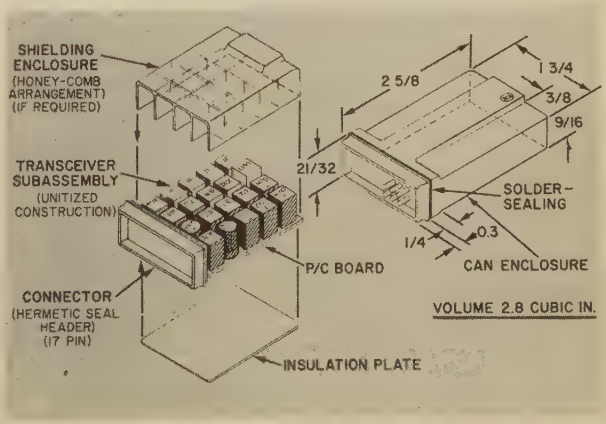


Fig. 7 - Transmitter-Receiver (Unitized Construction)

"unitized" construction of the Transmitter-Receiver Figure 7, features a single printed circuit board for interconnecting the modules, special transistors and associated heat sinks. This construction also provides a suitable anchor for a special 17-pin hermetically sealed connector. The design permits installation of an internal honey-combed enclosure which provides electrical shielding and mechanical support. The assembly slides into a light-weight metal enclosure which is then solder sealed.

The incorporation of the complete FM Radio Set including batteries, antenna, microphone, and control box, as part of the infantryman's combat helmet is obviously the next step.

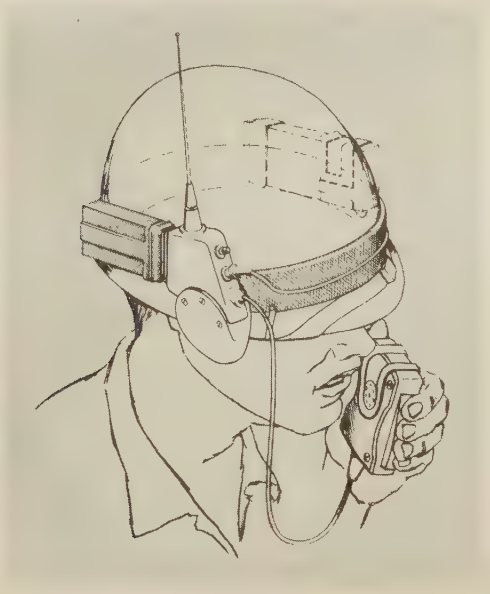


Fig. 8 - Combat Helmet Transmitter-Receiver

The communication range of the Radio Set, Figure 8, is about ½ mile and it is designed for operation on any one of 12 discrete crystal-controlled channels in the range of 38 to 51 megacycles. The Radio Set, assembled on an elastic band fitting the standard helmet, weighs 20 ounces. The control box mounted near the ear includes an earphone subassembly which is adaptable for either right or left hand mounting to meet the individuals physiological needs or preference.

We now discuss, the URC-11, used by the AF as a Personal Rescue Receiver-Transmitter. The present tube-type equipment weighs 5 pounds. The Micro-Module version, Figure 9, reduces the physical volume of this equipment to approximately 12 cubic inches, of which only five cubic inches is taken up by electronic circuitry. The design, as proposed, will be a self-contained, self-powered radio set capable of transmitting on a single predetermined frequency in the UHF band between 240 and 260 megacycles. The receiver will operate on a single preset frequency in the same UHF band. The expected range is about 2 miles. This radio set is contained in a watertight, aluminum

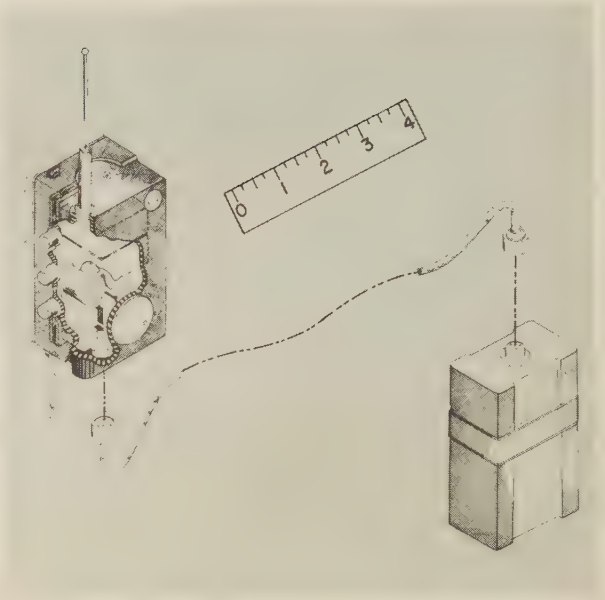


Fig. 9 - Personal Rescue Set

case and is small enough to be comfortably supported in one hand. The estimated weight, exclusive of battery and cable, is approximately one and a half pounds. The control functions are: TONE, TRANSMIT, RECEIVE.

Each push-button switch is of the multiple-contact, single-throw, normally off variety. A locking device permits the operator to mechanically maintain any one or all buttons in the "OFF" or "ON" position. The antenna, which is a telescoping, quarter wave-length, whip-type has an extended length of approximately 15 inches. When not in use, it collapses into a recess in the case. The battery, contained similar-sized pack, consists of mercury-type cells.

A number of alternate methods of repackaging the Personal Rescue Radio Set could be considered. One possible configuration, Figure 10, would be in the form of a shoulder pack or harness design as shown. One surface of this unit would be contoured to fit the shoulder and strapped firmly in place. The combined microphone-speaker would be in close proximity to both the ear and the mouth without requiring the use of hands. The antenna, similar to a metal tape ruler, would be released and erected automatically by a pull chain mechanism.

Although UHF operation is provided with the above configuration, VHF operation could be added. This would require an additional super-regenerative receiver circuit and an additional single stage power amplifier to cover the VHF band (120 to 130 megacycles). In addition, necessary switching to change from UHF to VHF would be needed.

Attention is now directed to the AN/PRC-25 Portable Radio Set, shown with its Micro-Module equivalent, Figure 11. It is designed to operate on any one of 920 instantly available crystal controlled channels spaced at 50 KC intervals over a frequency range of 30 to 75.9 megacycles. The

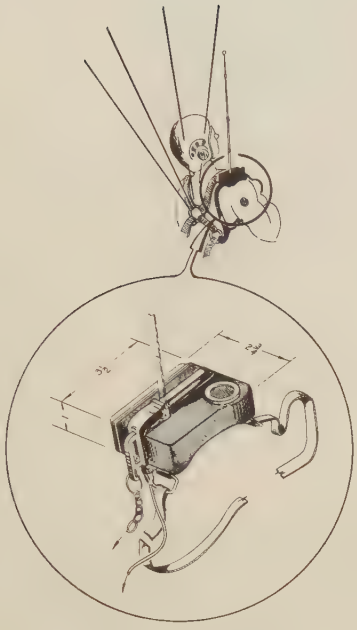


Fig. 10 - Shoulder Pack Radio-Set

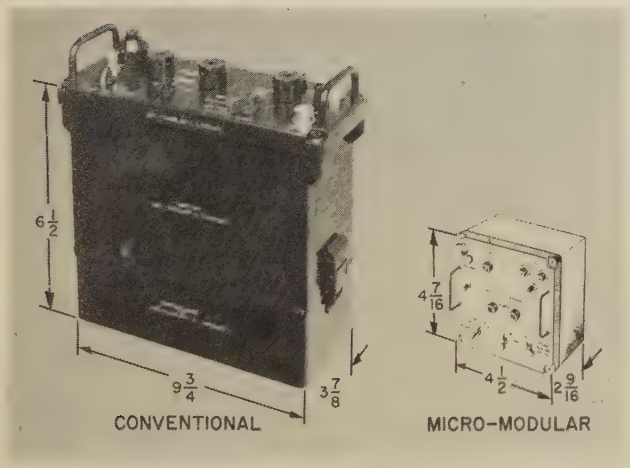


Fig. 11 - AN/PRC-25 Portable Radio Set

Micro-Module version includes a control panel and tuning mechanism, a chassis housing, an enclosure case, and plug-in type subassemblies. Approximately 46 modules containing about 510 components are used. This construction represents a size reduction of about 5 to 1 over the conventional equipment, exclusive of battery compartments.

Through use of Micro-Modules and the new techniques, we may develop a Single Sideband Communications Receiver, Figure 12.

The receiver occupies 300 cubic inches and weighs approximately 10 pounds. The design could include provisions for 22 plug-in units and space for 88 Micro-Modules. Compared to the requirement for a present-day-tube operated communication receiver performing the same function, the estimated over-all size reduction is 5 to 1. However, reduction of 10 to 1 is obtained in the volume of the electronic portion of the same receiver- which is really significant. This receiver would feature Micro-Module construction - incorporating the latest advanced circuitry with special electrical tuning techniques. The electrical tuning of the frequency control reduces the mechanical complexity by the elimination of critical gearing and cam-differentials. Compared to a transistor circuit performing the same function a tunnel diode will permit reduction of circuit components by 2 to 1 or better. Other advantages of advanced techniques are:

(1) Relatively low cost;
(2) Not effected by high humidity - requires no hermetic sealing;
(3) Low noise device.

The HF Single Sideband Communications Receiver design as described, represents a realistic step toward complete integrated circuitry. This receiver tunes a frequency range of 2 to 30 megacycles. The proper bands are automatically selected while the frequency controls are being set. The important characteristics of this receiver are:

(1) High Long Term Stability of 1×10^{-8} ppm
(2) Extreme Operational Simplicity (no cams, gears of differentials)
(3) Long Term Reliability - Part failure rate of 0.13% per 1000 hours
(4) High Density Micro-Module Packaging - Minimum size and weight
(5) Extremely Low Power Requirement

We have seen how Micro-Module packaging techniques have been proposed and applied to military equipments. Let us now project our attention to non-military or commercial applications.

Miniaturization played an important part in the success of Project Score which marked a significant first in electromagnetic communication payloads. The "Talking Atlas" gave the first hardware demonstration of the long discussed satellite radio-relay system, and also marked the first practical explanation of space research to non-military pursuits.

To illustrate the Micro-Module application, the FM receiver and the electronic control unit of the system were selected for consideration.

Figure 13 shows the packaging arrangement of the FM receiver and its equivalent Micro-Modular version. Specific circuit functions are integrated in 12 Micro-Modules, containing such micro-elements as resistors, transistors, capacitors, and diodes. A resulting volume reduction of 3 to 1

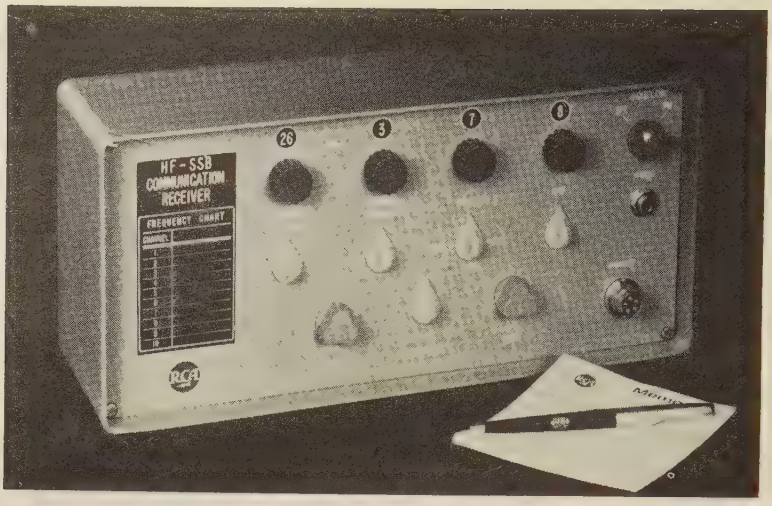


Fig. 12 - HF Single Sideband Receiver

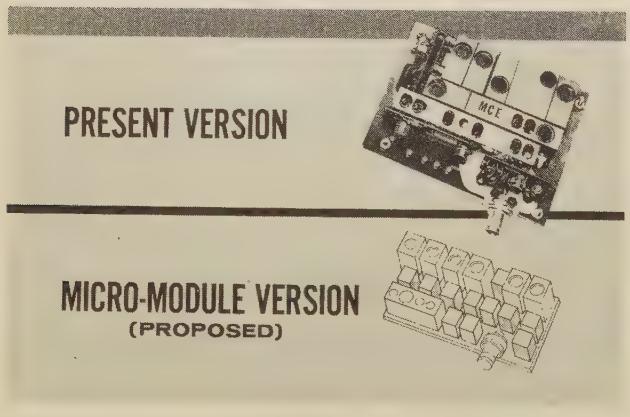


Fig. 13 - Command Receiver

Fig. 15 - FM Pocket Receiver

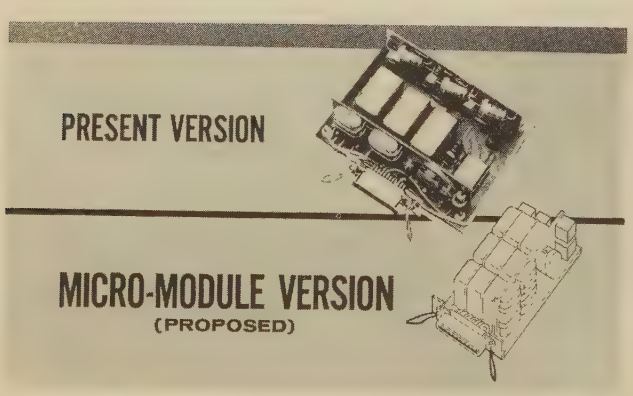


Fig. 14 - Electronic Control Unit

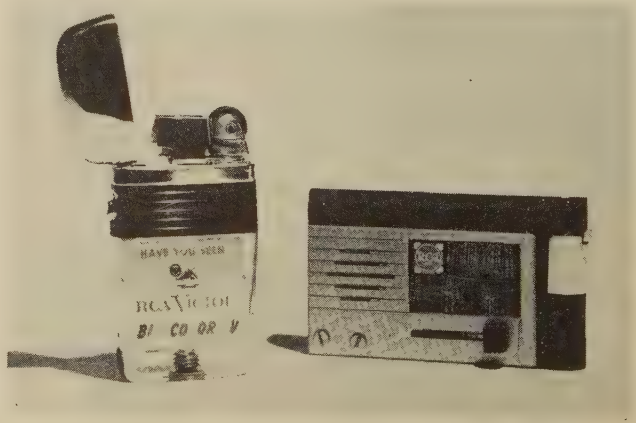


Fig. 16 - Micro-Module Broadcast Receiver

with an accompanying weight reduction of 5 to 3 may be obtained. Similarly, Figure 14 shows the arrangement of the electronic control unit and its Micro-Module counterpart. In this case, circuit functions are integrated into four Micro-Modules. The volume and weight reductions are similar to those stated above for the FM receiver.

In terms of total missile payload reduction the weight saving is 18 ounces. By circuit changes it is possible to further reduce this volume and weight without adversely affecting performance. It can be seen that other electronic portions of the payload may also be micro-modularized for further weight reduction. The biggest advantage, however, is increased reliability through redundancy. The impact here holds great promise for more reliable missiles and satellites by permitting circuit redundance at expense of very little extra equipment volume.

In 1957 a 150-megacycle micro-miniature FM pocket receiver, completely self contained from antenna to loudspeaker, was first marketed. This 20-cubic inch paging receiver, Figure 15, was developed and is currently produced by RCA Commercial Electronic Products. The New York City Police Department used quite a number of these receivers in a successful field test for receiving instructions from a single fixed police F-M transmitting station. These receivers have been used at the Indianapolis Speedway by camera crews. In addition installations have been made by RCA for DuPont, San Diego Naval Hospital and others. Even by present standards this represents an excellent attainment in miniaturized design using conventional components. Since the advent of Micro-Modules we have come a long way. The Micro-Module Broadcast Radio Receiver is an experimental model and is approximately the size of a cigarette lighter. This receiver, Figure 16, occupies only a volume of 2 5/8 cubic inches. It is a complete superheterodyne receiver containing a converter, two stages of IF amplification, a detector, and two stages of audio amplification. The circuitry is contained in six Micro-Modules, all of which can be seen through the transparent window of the set. This receiver contains 44 circuit elements, including six transistors. It also contains a slide tuning capacitor (developed especially for this set), a suitable speaker, a ferrite antenna, and mercury type battery cells.

You may now, want to ask, "A Broadcast Receiver?" How soon and how much? This question may best be answered as follows.

Current Micro-Module manufacturing is accomplished through pilot-run production methods which, obviously, is too costly in comparison to present retail price levels. The disciplined geometry of micro-elements and Micro-Modules make them highly suitable for eventually complete mechanized manufacture. Such mechanization leads directly to a high standard of quality at low cost. Within the next two years quantity manufacture could produce a Micro-Module receiver whose cost would be at least as low as that of its conventional counterpart.

As an outgrowth of the Combat Helmet as discussed earlier, a minified radio-transmitter set can be effectively employed as part of a fire department communications system for municipal, commercial airports, as well as military aircraft landing or carrier operation. Specifically, the set can be used as part of the helmet of a fire fighter's protective suit. In such an application, a complete system encompasses a receiver, transmitter, battery supply and antenna and microphone. It is the design goal to have this type radio set operate satisfactorily over a range of 500 yards under normal conditions. The transmitter and the receiver will operate on separate frequencies, 49.8 mc and 49.4 mc respectively, in order to eliminate man to man communications and give the fire chief access to all his subordinates, at all times. With normal signal input it is anticipated that the receiver will deliver the adequate amount of 5 to 10 milliwatts of output power to the earphone. The battery should supply about four hours of continuous operation. The design in each case consists of 17 basic circuits constituting 20 Micro-Modules.

In one of the constructional concepts, which has been considered, the Figure 17, the receiver and transmitter are included in the helmet in separate cases. Each unit is complete with its own battery and accessories. The receiver assembly occupies a volume of 4.75 cubic inches and weighs 9 ounces; the transmitter is of similar volume but weighs 12 ounces. An added feature of this design is that the two units can be combined to form a general purpose or "twin-pak" communication set.

A more powerful version of the fire fighter's gear is adaptable to police work or special investigations where emphasis is placed on portable light-weight equipment which can be camaflouged or easily concealed. Such a transceiver may also serve a Civil Defense Communication System supplying two-way voice operation within a 10 mile range

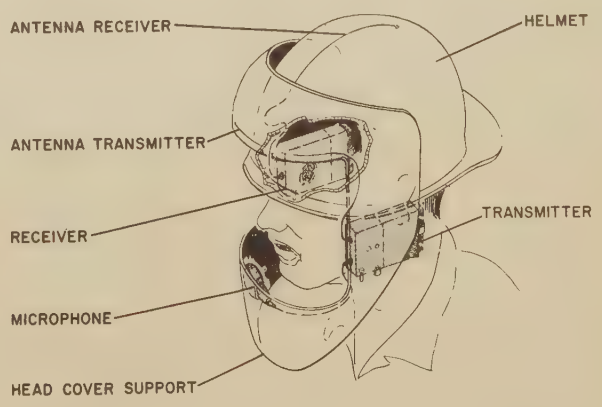


Fig. 17 - Fire Fighters "Twin Pak" Radio Set

between various personnel and a central control station.

When the Micro-Module Concept is applied to the "Citizens Band" Radio, the two-way Radio Telephone Communication System Unit, smaller than a king-size cigarette pack, soon may be available. Especially since the opening of the Class D - 27 megacycle "Citizens Band", radio telephones for outboard motor boats and small craft usage, is economically possible. In time, a Micro-Module Radio-Telephone should be sufficiently economical to come within the price range of general public demand.

Similarly, such a compact communication unit would find other numerous applications, in the home, on the farm, on theater stages, and at sporting events. In the construction of large buildings, bridges, highways, and telephone line erection, the Micro-Modular transceiver, as part of the worker's apparel or personal equipment, will facilitate the tasks with increased safety.

The wall type television set, its surface determined by the expanse of the viewing tube and the rest of the components in the rim of the picture frame, is nearer to reality. Still looking ahead, RCA contemplates record players, dictation machines, as well as tape recorders that could fit in a pocket or small purse.

CONCLUSION

It has been pointed out how the Micro-Module has been and can be integrated effectively into various military and commercial communications equipments. This technical break-through is beginning to embrace the communications industry and its hardware. Production capability should be possible by 1961. True, the aim is ultra-miniaturization, however, making something smaller alone is obviously NOT enough. The building block characteristic of the Micro-Module, with its standardized structure, its circuit function application and its mass production capability is geared to:

(1) Obtain greater reliability;
(2) Ease the production effort at reduced cost;
(3) Provide simplified maintenance - by focusing design and production trends in the direction of disposal at failure maintenance philosophies.

Also it has been shown that the Micro-Module Concept permits inclusion of solid state devices in combination with conventional micro-elements as fast as the development of these new devices permits. Particularly the tunnel diode which may be used as an IF amplifier in the High Frequency Single Sideband Communications Receiver - projects an example of practical design feasibility. In addition progress is beginning to be made with other advanced component developments such as integrated unipolar transistors, mesa transistors, parametric amplifiers and others, which lie within the immediate scope of the Micro-Module Program. These startling innovations in basic circuitry are the miniaturized building blocks of tomorrow's equipment. They will make it possible to produce vastly improved communications equipment for national defense and space explorations, as well as for the consumer and industrial markets.

THE APPLICATION OF SEMICONDUCTORS IN A 860 MC RADIO RECEIVER

L. G. Schimpf
Bell Telephone Laboratories, Incorporated
Murray Hill, New Jersey

Summary

An all solid state, FM radio receiver operating at 860 mc is described. The active elements used are transistors and variable reactance diodes. The receiver is of the double conversion type with IF frequencies at 63 and 107 mc.

The first local oscillator chain uses four stages to generate a signal at 923 mc. The crystal controlled oscillator and two frequency doublers use diffused base transistors. The output of the second doubler, at a frequency of approximately 300 mc, is then tripled in a circuit employing a varactor diode in order to obtain the desired local oscillator signal.

The receiver has a 6 db bandwidth of 30 kc. If the output is equalized for a phase modulated signal, 10 db of noise quieting is obtained with a 0.8 microvolt input signal.

Introduction

A new type of mobile telephone system is now under investigation in the research department of the Bell Telephone Laboratories. More detailed information on this system will be given in a group of papers which will follow in another session. For our purposes, it will suffice to say that the proposed system will be multichannel with a number of these channels available to each subscriber. Because of this multichannel feature, the required frequency spectrum must be fairly wide. This being the case, it is thought that such a system might be operated on frequencies as high as 750 or 900 mc. Channel selection, ringing, supervision and other functions associated with normal telephone operation, would be under control of the central office over the radio link. This information would be sent as a coded signal to tell the mobile unit what to do. Since this information is very important, it must be received with little chance of error. In order to improve the accuracy, several methods or combinations of these methods may be used. These include error checking and correcting codes, transmitting the same information more than once or transmitting the information over more than one channel. The use of more than one channel looks attractive since it results in frequency diversity. This requires that the receiver must have at least three channels, two for transmission of coded information or "data" and one for voice communication. Additional channels may be required for other functions.

Since the frequency range is quite a bit higher than used for other mobile telephone service, it is necessary to get data on propagation, fading, shadow loss and noise. In order to make this investigation, one of the UHF television channels (866-872 mc) is being used on an experimental basis in a local area. These tests should be performed with a receiver located in a moving, mobile unit. A receiver that is small, light and economical in power consumption is desirable. Such requirements go hand in hand with semiconductors. Since most future mobile telephone systems will use semiconductor receivers, it was decided to investigate the design of such receivers for this frequency band. This paper will describe an early experimental model of such a receiver.

Design Considerations

With this as a starting point, a receiver can be blocked out as shown on Fig. 1. The selection of the IF frequencies was determined by the required selectivity and attenuation of spurious responses rather than the characteristics of commercially available semiconductors. With this procedure, it is then necessary to consider what semiconductor will best fit into any block in the receiver. A short discussion of the function of each circuit of the receiver will follow. After that, each block will be considered in more detail in order to show how the circuit may be synthesized.

Fig. 1 shows the antenna connected to the first mixer through a passive, preselection network without RF amplification. Radio frequency gain with low noise presents some difficulties at these frequencies with either semiconductors or vacuum tubes. The use of an amplifier with a poor noise figure, may result in a receiver with a poorer signal-to-noise ratio than if one were to go directly into the converter. It is true that a radio receiver employing an RF amplifier using a low noise device with sufficient gain to overcome the noise generated in the mixer, will result in a receiver which will detect lower level input

signals than the setup shown on Fig. 1. In many locations, this increased sensitivity may be useless due to interference from noise sources in the vicinity of the antenna. Since we did not have a great amount of data available on the magnitude of the ambient noise, it was decided to proceed without RF amplification. From field data obtained with this receiver, we will be able to determine if the ambient noise situation is such as to warrant a low noise stage of radio frequency amplification.

If, as a result of field studies, it is decided that a stage of RF amplification is necessary, several possibilities present themselves. At the present time, commercially available transistors would not have sufficient gain or low enough noise figure for this service. Some experimental units could be used and will no doubt be available on a limited commercial basis in the near future. Varactor diodes, to be discussed later, and which are available on a commercial basis look attractive as an all solid state amplifier for this frequency band.

In the interest of simplicity, the frequency values shown on Fig. 1 have been rounded off. The working receiver is only a few megacycles removed from the values indicated. In addition to the 860 mc input from the antenna, the output of the first beating oscillator is also coupled to the mixer.

The difference frequency is amplified by a 63 mc transistor amplifier with a 6 mc bandwidth. This two stage amplifier has a gain of about 35 db. A second mixer follows. A 10.7 mc output signal is derived in this stage, this frequency being the difference between the second local oscillator and the 63 mc IF. At this point the major portion of the receiver selectivity is attained. As shown on the block diagram, the received signal is split into several paths at this point. In the experimental receiver, three channels are used, two of these are about 30 kc wide, separated by about 1 mc in order to study the effects of frequency diversity on data transmission. Crystal filters are used to separate these channels. The third channel is used to measure and check frequency stabilization methods. One such method is indicated on Fig. 1. A 200 kc wide channel at 10.7 mc is selected by a four pole, LC network. This filter is followed by a 10.7 mc IF amp and limiter. A frequency discriminator, employing quartz crystals measures the input frequency. If the local oscillator is off frequency, a dc output will be developed by the discriminator. This signal is then amplified and sent back to the first local oscillator, causing a shift in its frequency so as to reduce this error. It may also be possible to build the required stability into the first local oscillator without too much complication.

The various parts of the receiver will now be examined in more detail starting with what is usually termed the "front end" and proceeding toward the signal output circuit. The front end consists of the first local oscillator, preselector and first mixer.

Before going ahead with a description of the first local oscillator, it might be well to digress a moment in order to discuss the properties of semiconductor devices needed in the local oscillator chain. At frequencies up to 300 or 400 mc, diffused base transistors have reasonable gain and power handling capabilities. At 923 mc, the output frequency of the first local oscillator, only special and selected transistors would be capable of delivering the few milliwatts of power required to drive the first mixer. A semiconductor device which looks very attractive for application in this frequency range is the "varactor". This device is a diode of special construction, designed to have small losses in the frequency range where it will be used. In the 1000 mc region, the diode is of small cross-sectional area and made from silicon. Such a diode, when biased in the nonconducting direction can be represented to a first approximation by the network shown on the left of Fig. 2. The capacitor, C_j, is the junction capacitance with a value of a few micro-microfarads. R_p has a high value when the diode is biased in the back direction and in the frequency range we are considering, can be neglected. R_s is the series resistance of the junction, which by proper fabrication can be made to be about one or two ohms. Since this is the only dissipative element, the back-biased diode can be considered as a capacitor with a fairly high value of Q. Values of 100 are not uncommon at 1000 mc. The value of this capacitance is a function of the voltage across it. A typical case is shown on Fig. 3. (The negative voltage indicates bias in the back direction.) Most diodes that are useful in this frequency range are limited to a back bias of 8 or 10 volts due to reverse breakdown. The diode can be biased a few tenths of a volt in the forward direction and still maintain its high value of Q. This variation of capacitance is the parameter that gives the device several useful properties at high frequencies.

If a varactor is connected in the circuit shown on Fig. 4, the capacitance will vary during the cycle of the applied voltage. As a result, the wave form of current will be distorted causing the generation of harmonics. By tuning L_1 C_1 to the desired harmonic, power at this frequency can be delivered to the load. Since the loss in the diode is quite low, the efficiency can be quite high. As an example, the third harmonic can be generated with a loss of less than 6 db. The diode is used in this manner in the local oscillator chain. The variable capacity diode may also be used as the active element in high frequency, low noise amplifiers, mixers and switches.

Circuit Description

The stages of the oscillator chain are shown on Fig. 5. The oscillator uses a diffused base

transistor in the common base connection. Oscillations are maintained by placing a series resonant quartz crystal in a low impedance feedback path between the output and the input of the amplifier. This is followed by two doubler stages using the same type transistors. The last doubler delivers $20 \cdot 10^{-3}$ watts at 307.6 mc. This stage feeds a varactor diode which is set up in a circuit to triple the input frequency. A tuned resonator was used to insure that the output circuit operates at a low loss. The input loop is adjustable so that the load impedance presented to the diode can be optimized for efficient harmonic generation. The dc bias is also adjustable for the same reason. This circuit will deliver a power of $5 \cdot 10^{-3}$ watts at 923 mc which is sufficient to operate a diode mixer.

A simplified schematic of the oscillator is shown in Fig. 6. The output circuit is tuned for parallel resonance at a frequency of about 76.9 mc. This will present a high impedance to the collector of the transistor and the stage will have gain from the emitter to the collector. The crystal is used at its fifth overtone and completes the feedback path at a frequency corresponding to this overtone. Small frequency corrections can be made by the variable capacitor, C_1. The doubler stages are quite similar excepting that the feedback path does not exist and the output circuit is tuned to twice the frequency of the input signal.

The connections between the oscillator chain and the mixer are indicated on Fig. 7. Since the oscillator output is only a few milliwatts, the use of a high loss coupling circuit so as not to shunt the input signal path is prohibited. If one uses a $\frac{\lambda}{4}$ line between the oscillator resonator and its junction with the mixer line, a high shunt impedance is presented to the signal at this point due to the short circuit at the resonator. Efficient oscillator coupling is thus obtained without shunting the signal. A stub tuner is provided to obtain efficient transfer of power from the antenna to the preselector which consists of two resonators coupled through an aperture. This is equivalent to a high Q, double tuned circuit. A double stub tuner is used to match the output to the mixer. The tuned circuit at the output of the mixer serves two purposes; first, provides a low resistance dc return for the mixer and second, presents the proper impedance at the IF frequency.

The first IF amplifier employs two stages in a common base connection. The circuit of a typical stage is shown in Fig. 8. The gain vs frequency curve of the two stage amplifier is plotted in Fig. 9. It will be noted that the amplifier has a bandwidth of about 6 mc. Since the receiver front end is at least this wide, the receiver is broadband through this IF stage.

The second local oscillator is similar to the crystal controlled oscillator previously described. The second mixer uses a transistor in a circuit quite similar to an amplifier stage. The local oscillator and input signal are mixed in the base-emitter circuit while the collector is tuned to the difference frequency. A functional representation of this circuit is shown in Fig. 10. At the output of the mixer, the main part of the selectivity of the receiver takes place. Also, at this point, the various channels are separated.

Each channel is then connected to a 4 stage, IF amplifier. A typical stage is shown in Fig. 11. Since the filters ahead of the amplifier provide the major portion of the selectivity, these stages are fairly broad with a loaded Q of about 10. Since these amplifiers are also used as limiters, a third winding on the output transformer is coupled to two diodes to clip the output signal at a level a little lower than that at which the following transistor stage would overload. The level at which clipping occurs is determined by the turns ratio on the transformer. This circuit is applied to three stages of the amplifier. Each amplifier is terminated in a crystal discriminator to convert the FM signal to an amplitude variable output.

Measured Characteristics

Measurements have been made in order to compare the performance of this receiver with others used in mobile telephone service. This comparison was made by equalizing the audio output for detection of phase modulated signals and measuring the audio output as a function of the radio frequency input. This data was obtained for inputs of both 63 mc and 860 mc. The data for 63 mc is shown in Fig. 12. It will be noted that full limiting occurs at an input of about 0.6 microvolts. At this input, the receiver quieting is about 16 db resulting in a 28 db signal-to-noise ratio. This is about the same as was obtained with vacuum tube receivers having a 10 db noise figure.

Since this receiver does not employ a RF amplifier, the performance will be degraded at the RF input frequency due to the noise and loss in the mixer. A noise quieting curve for an input frequency of 860 mc is shown on Fig. 13. If this curve is compared with the corresponding curve on Fig. 12, it will be noted that the mixer introduced a net loss of 8 db. A signal output curve was not taken because of the lack of an FM signal source at this frequency.

The selectivity of the receiver is determined by the crystal filters and is similar to that obtained on single conversion receivers employing such filters in 10.7 mc, IF amplifiers. The 6 db bandwidth of the filter is 30 kc. The main image, 126 mc away, is down by 80 db. A number of spurious responses exist and no attempt will be made to list them or explain their origin. The worst of these was measured at 60 db below the desired input.

Acknowledgments

The author wishes to express his gratitude to R. M. Ryder and A. E. Bakanowski, who not only made available the varactor diodes but were frequently consulted concerning their circuit applications. He is also indebted to M. F. Wazowicz for taking care of the many details of construction and for measuring the performance of the receiver.

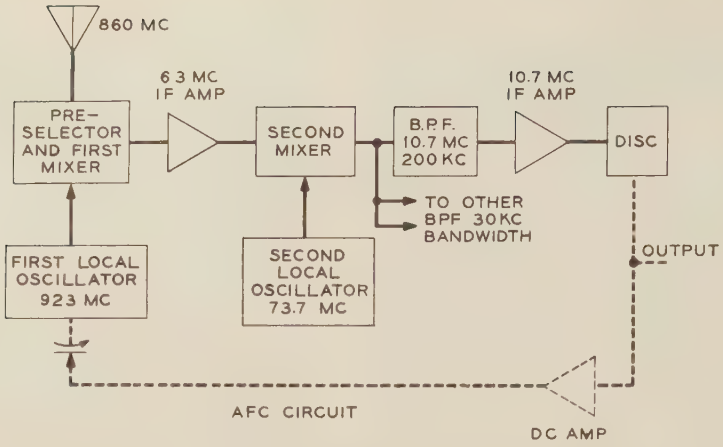


Fig. 1 - Block Diagram of the Receiver

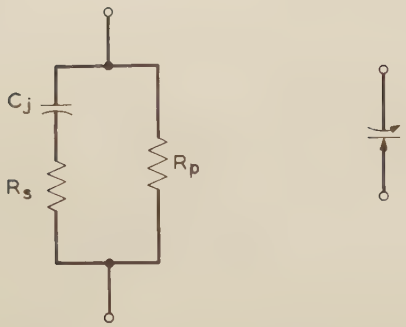


Fig. 2 - The High Frequency Equivalent Circuit of a Varactor Diode is Shown on the Left.

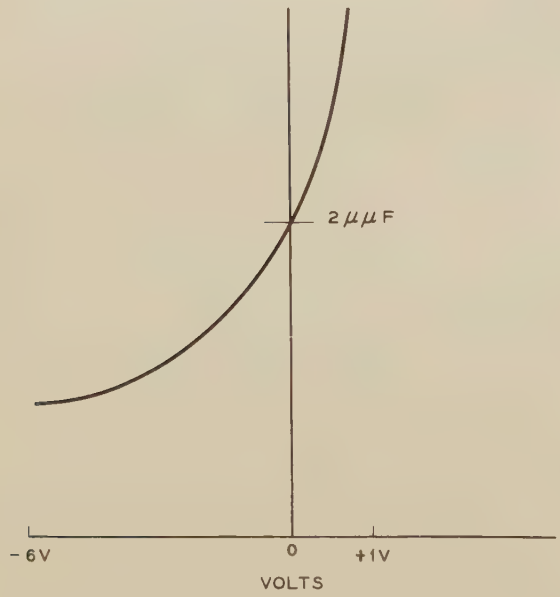


Fig. 3 - The Capacitance of the Diode is a Function of the Applied Voltage

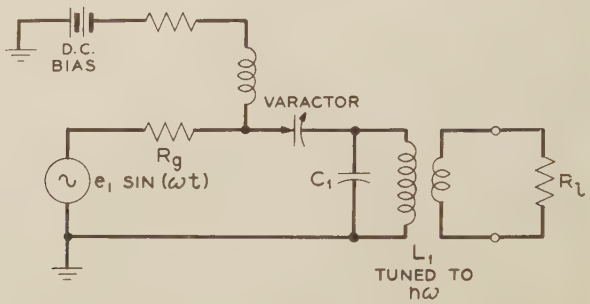


Fig. 4 - Varactor Diode in a Circuit for Harmonic Generation

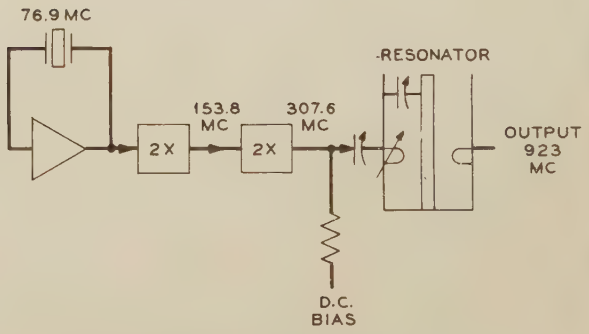


Fig. 5 - Functional Diagram of the First Local Oscillator

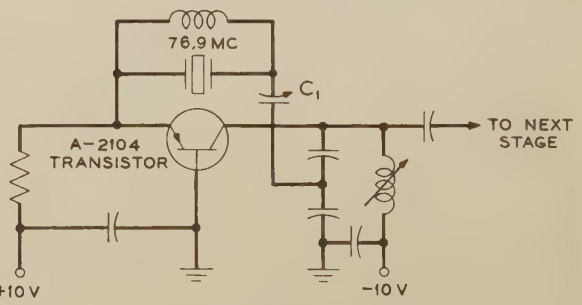


Fig. 6 - Simplified Schematic of the First Local Oscillator

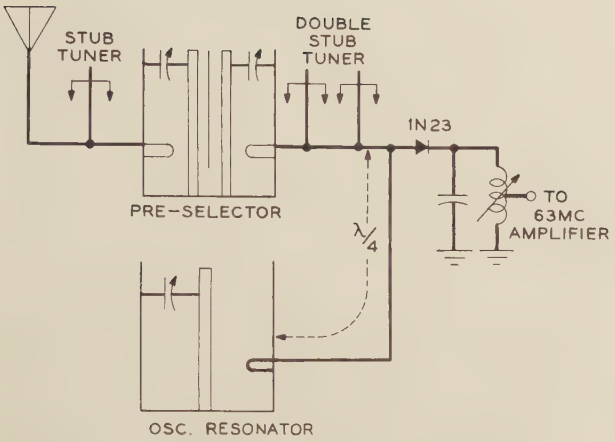


Fig. 7 - Functional Diagram of the Front End of the Receiver

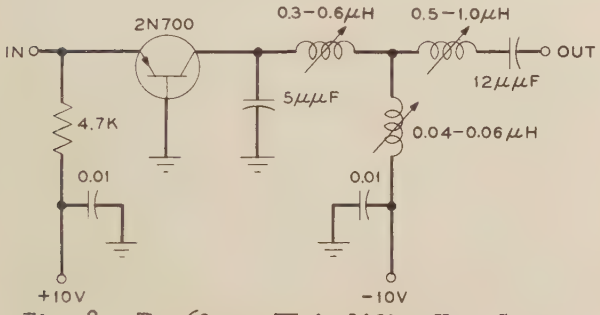


Fig. 8 - The 63 mc, IF Amplifier Uses Common Base Stages.

Fig. 12 - Noise Quieting Curve for a Signal Applied to the Input Terminals of the 63 mc, IF Amplifier. (Photo at right)

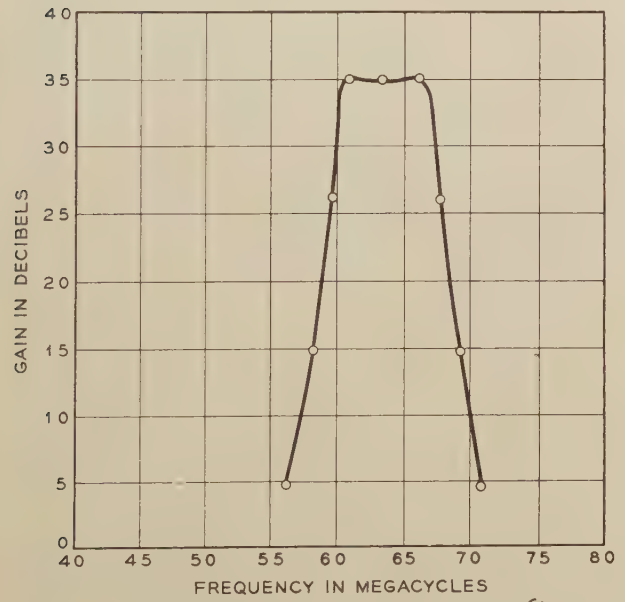


Fig. 9 - The Gain Characteristic of the 63 mc, IF Amplifier

Fig. 13 - Noise Quieting Curve for a Signal Applied to the Input of the Receiver. (Photo at right)

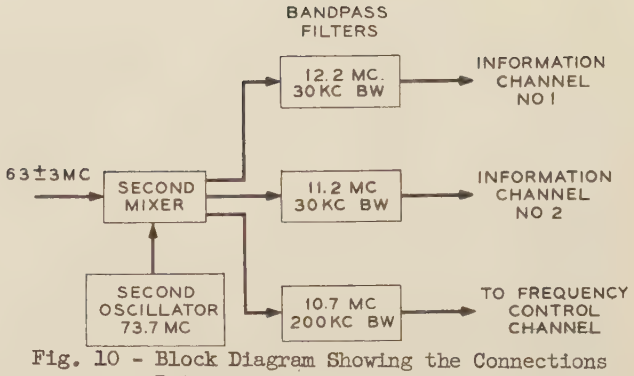


Fig. 10 - Block Diagram Showing the Connections Between the Second Mixer and the Bandpass Filters

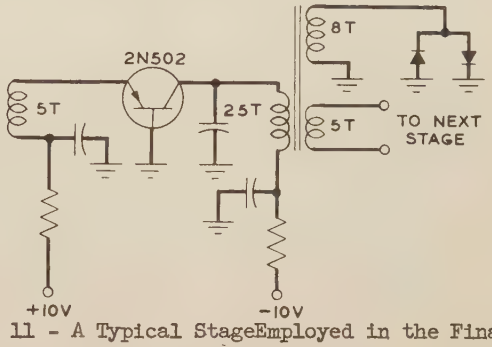
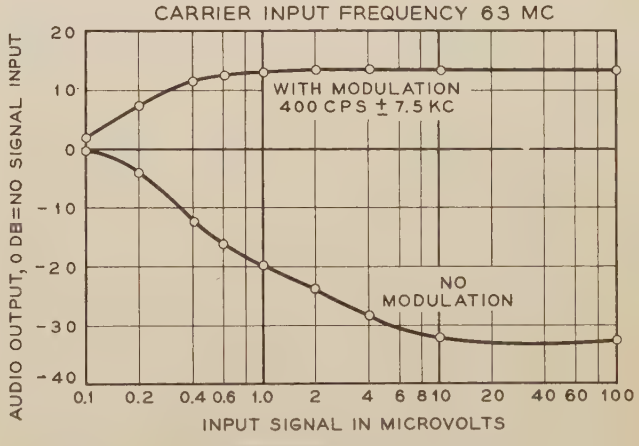
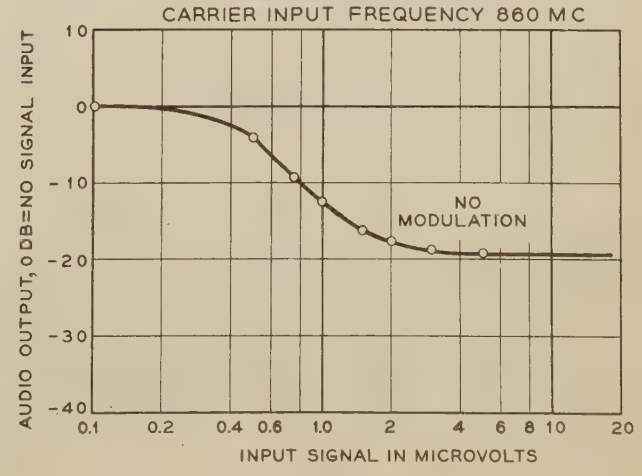


Fig. 11 - A Typical StageEmployed in the Final IF Amplifiers


CARRIER INPUT FREQUENCY 63 MC
WITH MODULATION 400 CPS ± 7.5 KC
NO MODULATION
AUDIO OUTPUT, 0 DB=NO SIGNAL INPUT
INPUT SIGNAL IN MICROVOLTS



A HYBRID MOBILE TWO-WAY RADIO

R. A. Beers, Jr.
Radio Corporation of America
Camden 2, N.J.

Summary

Maximum effective use is made of transistors and tubes to achieve a mobile unit with superior performance characteristics and considerably improved realiability without, at the same time, appreciably increasing the cost of the radio equipment. No aspects of performance have been compromised to accomodate the use of transistors.

A rather unique mechanical design achieves a high degree of heat transfer from the interior of the mobile unit to the surrounding outside air. Both maintenance and production have been simplified by judicious use of "Circuit View" printed wiring in the more complex portions of the equipment. In addition, all critical adjustments have been designed out to further simplify routine maintenance adjustments. Maintenance may be performed using conventional test equipment and techniques.

As the state of the electronic art progresses incessantly, the task of the designer of vehicular communications equipment becomes in certain respects, less difficult, and in other respects increasingly more difficult. The task becomes less difficult because of an ever increasing supply of knowledge, materials and components from which to choose. It becomes more difficult because many additional methods, materials, components, and circuits must be thoroughly evaluated in order that the best combination may be chosen to give the end user the most communications over the longest possible time, per unit of cost; commensurate with the present state of the art.

The following considerations are of paramount importance in the design of any piece of mobile communications equipment regardless of what combination of methods, ideas and components are selected:

1 - Reliability of Equipment
2 - Reliability of Communications
3 - Ease of Maintenance
4 - Battery drain
5 - Cost

Now, stated broadly; reliability of equipment means that the equipment should continue to operate over acceptable periods of time without deterioration of initial performance or without failure.

Reliability of communications means that the equipment should be able to perform adequately under all the strict operational requirements of the mobile communications services; such as, maximum spectrum utilization with its problems of intermodulation, desensitization, adjacent channel interference, spurious response attenuation, and co-channel usage.

The equipment should be so designed that it may be adjusted and maintained, using for the most part, existing skills, instruments and techniques.

The battery drain should be as little as practical. It should be sufficiently low to insure that the receiver may "stand-by" for the receipt of messages over long periods of time without the necessity of running the engine to keep the battery charged where operational requirements so dictate.

Any new equipment design must provide substantial improvements in one or more of these aspects without appreciably increasing the cost of the equipment.

The transistor, which by now has matured for ten years since it was first discovered, is one of the most interesting, most sensational and perhaps offers the greatest potential of any of the new devices. Knowing the exacting requirements; the overall sophisticated equipment design necessary to fulfill the needs of the mobile communications user; how best can the many advantages of the transistor be used to entirely fulfill all of these requirements.

Since 1953, we of the RCA Commercial Communications Products Department, have developed various models of transistorized land-mobile equipment. They have been thoroughly evaluated in the laboratory and in the field under actual operating conditions. One of these products, a completely transistorized pocket VHF receiver, was first placed on the market over two years ago. An improved modelwith vastly superior performance will be introduced in the near future.

In order to meet the strict technical performance requirements of intermodulation rejection, desensitization by strong signals nearby in frequency, and to obtain the required 100 db. of spurious and image response attenuation by practical means; while at the same time being able to maintain initial receivers sensitivity over long periods of time, would have required transistors that do not yet exist in commercial duplicatable quantities and at reasonable cost. So, in order not to compromise with the performance which the effective use of the land-mobile bands

dictate, the use of both transistors and tubes in the receiver were found to be necessary.

The result is a new mobile equipment with every performance characteristic equal to or better than its all tube predecessor. Transistors, a total of 13 or 15 of them depending upon RF power output, have been used where they will do the best job, on the other hand, tubes are used where they will do the best job.

To make installation and maintenance easier, the transmitter, power supply and receiver chassis are contained within a single housing (figure 1). To gain access to the chassis it is only necessary to turn the lock about 30° and pull the cover off, either straight out, or by lifting it up. The entire T-R unit can also be removed as easily from its mounting by turning the lock 30° and either withdrawing it straight out, or by lifting it directly up. (figure 2). When withdrawn from its mounting, the chassis assemblies are protected from accidental damage by the rugged malleable aluminum "wrap-around" heat sink.

The trunk mount unit can be readily converted to "under dash mount" by merely plugging the standard control box into the front of the radio unit and adding an under dash mounting bracket. (figure 3)

To achieve the successful marriage of transistors and tubes within a single housing, some unusual mechanical design was necessary. In order to assure that the radio unit operates over the normal environmental range of -30°C to + 60°C ambient temperature with a transmitter duty cycle of 20%, the heat rise within the case had to be accurately controlled at all locations. With the exception of the front panel, the entire perimeter of the case is a very efficient heat sink with an effective area several times that of a normal surface. (figure 4). The portion of the transmitter which generates a lot of heat is tightly coupled to the heat sink on the right side. The power supply switching transistors are mounted on the rear heat sink. The heat rise within the transistorized portion of the receiver is maintained at 15°C or less by the left heat sink.

The heat rise in this section is restricted to a small amount by enclosing the transistorized section within an insulated shield. The shield itself is thermally, poorly couply to the receiver chassis, but thermally, efficiently coupled to the heat sink.

The entire receiver thereby remains a single assembly which can be readily handled in production or can be easily reached for routine maintenance.

The receiver transistors, with the exception of the audio power output transistors are mounted on a "road map" type of printed circuit board. This not only isolates the transistors from sources of heat, but at the same time considerably reduces component and wiring congestion on the wiring side of the receiver chassis. (fig. 5 and fig. 6). Note that all the circuit wiring is silk screen reproduced on the visible component side of the board so that all connections can be traced even more readily that they could be traced with conventional point to point wiring. Even more, each component symbol is applied adjacent to its position on the board. Even the transistor collector, emitter and base connections are labeled.

Now, it was earlier mentioned that transistors were used where transistors would perform most reliably, and that tubes were used where tubes would provide the best performance at the present state of the art. Lets refer to the block diagram (fig. 7). It will be noted that the I.F. amplifier, the audio amplifier and squelch circuits of the receiver are completely transistorized. In these circuits, it is possible to use transistors with a long history of reliable duplicatable characteristics from one transistor to another, and in circuits which have been temperature stabilized over the temperature range of -30°C to +85°C with not more than a few db. change in performance. All the transistors employed in these circuits have sufficient high voltage rating to insure that they will not be damaged by transient or high signal voltages or currents. Their mechanical construction is such that they will not be damaged by the constant vibration and shock to which mobile equipment is subject.

Furthermore, they are now very little more expensive than a tube. It is also of interest to note that only five transistors are used in the IF amplifier compared to six tube sections in its predecessor. IF gain is sufficient to provide a minimum of 40 db. of limiting on front end noise. Likewise, the audio amplifier and squelch circuitry use 6 transistors compared to 5 tubes sections in its all tube predecessor. But, here one additional transistor is used in the class B audio output stage with a 7 db. increase in audio power output. The low level audio amplifier stages use NPN transistors so that noise can be more effectively filtered out of the customary "positive hot" car battery supply. The squelch function is performed by a very effective transistor switch which in either "on" or "off" depending upon the adjustment of the squelch threshold control and the presence or absence of noise. All IF adjustments are fixed, with the exception of the discriminator, insuring long trouble free operation without the necessity of maintenance adjustments.

The front end of the receiver uses either five or six tubes depending upon frequency band. The 7000 series tubes, specifically designed for mobile equipment operation, are used throughout. Adequate front end gain is provided to maintain the original sensitivity over long periods of

time. This gain is so proportioned and front-end linearity has been maximized to achieve the greatest amount of protection from intermodulation response and freedom from desensitization by strong signals nearby in frequency. A total of 14 tuned circuits are used in the front end to assure a high degree of spurious and image response attenuation. In addition, a single lumped 12 coil (fixed tuned) IF selector provides the selectivity.

The transmitter uses seven or eight rugged mobile series tubes, depending upon RF power output. A phase modulator capable of $\pm$ 90° phase shift at the oscillator frequency with very low distortion enables either wide band or narrow band modulation with a frequency multiplication of only twelve times. Sufficient drive is made available at the final grids to assure maximum operational reliability under all conditions.

The power supply is fully transistorized. Either two or four switching transistors are used depending upon transmitter power. All rectifiers are silicon.

Space is provided in the power supply for optional "Quiet-Channel", tone operated squelch. This function can be supplied initially or readily added any time later. Installation is very simple. It consists of bolting a sub-chassis to the power supply chassis with four screws and plugging in a cable to the receiver and transmitter chassis. Resonant Reed Relays are used to both generate and respond to any of the 36 standard EIA tone squelch frequencies.

The use of a combination of transistors and tubes has resulted in some note-worthy improvement in performance and reliability. At the same time, not one single performance characteristic has been degraded to accomodate the use of transistors.

Battery Drain has been reduced, 33% with the switch turned to the "full on" position. It has been reduced to 2.2 amps on the "standby position" A full 60% reduction.

The audio output has been increased a full 7 db. to 5 watts with less than 10% distortion.

The equipment is available with full power output of 30 or 60 watts in the 150 mc band and 50 or 100 watts in the 25 - 50 mc band.

A typical 150 mc receiver has a sensitivity of less than 0.4 uv input for 20 db. of noise quieting. A typical 25 - 50 mc receiver requires less than 0.2 uv input for 20 db. of noise quieting. (figure 8)

All receiver spurious and image responses are attenuated more than 100 db. Intermodulation response due to a same side of carrier adjacent channel and alternate channel signal is down 65 db. or more. (figure 9).

A strong signal nearby in frequency must be 96 db. or more above the desired signal amplitude to desensitize the receiver 6 db. (figure 10). Adjacent channel selectivity is 85 db. or more by the two signal method of measurement, 100 db. plus by the noise quieting method. (figure 11)

Transmitter spurious radiation is attenuated 70 db-minimum out of band and 100 db. minimum within the band. System signal to noise ratio exceeds 40 db. on narrow band operation.

The ambient temperature range of operation is from -30°C to +60°C with a 20% transmitter duty cycle. The oscillator stability is within $\pm$.0005% over this temperature range using heated crystals.

In conclusion, it is apparent that the most practical use of the available state of the art tools have been lined up in the RCA LD-50 and LD-150 to create a two way mobile communications radio which achieves a new high in both reliability and the performance required for todays highly congested Land-Mobile service bands.

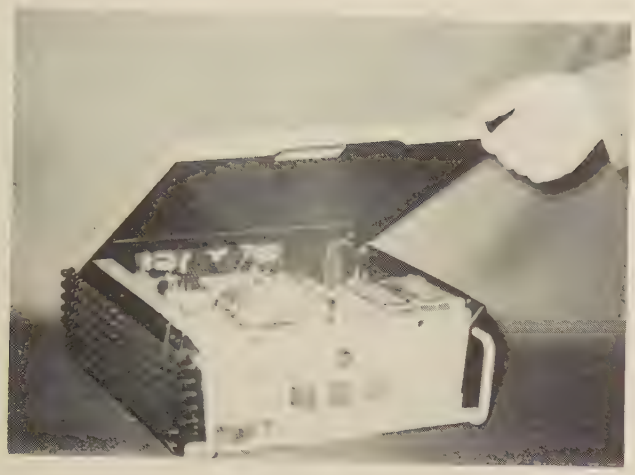


Fig. 1 - Mobile Unit in Single Housing

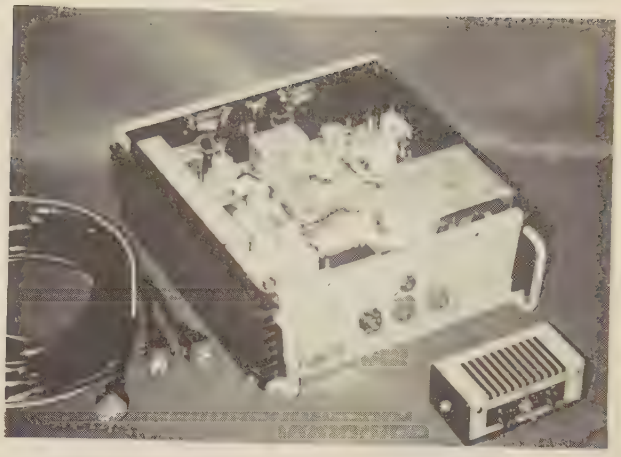


Fig. 4 - Heat Sink Forms Perimeter

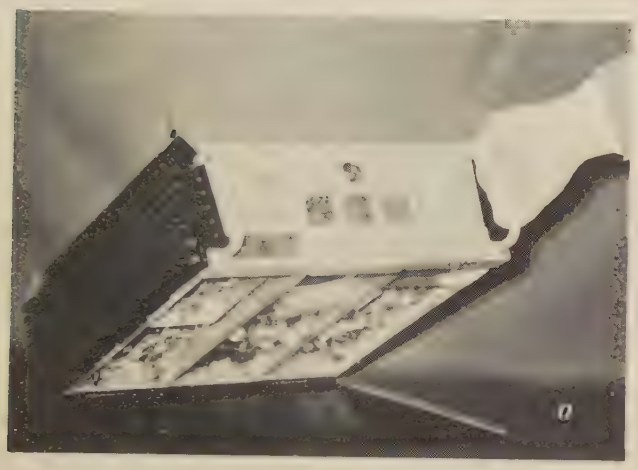


Fig. 2 - Removing Unit from Base Plate

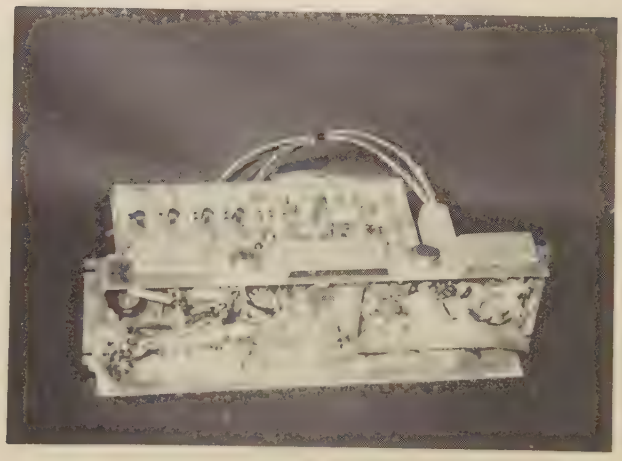


Fig. 5 - Receiver Chassis

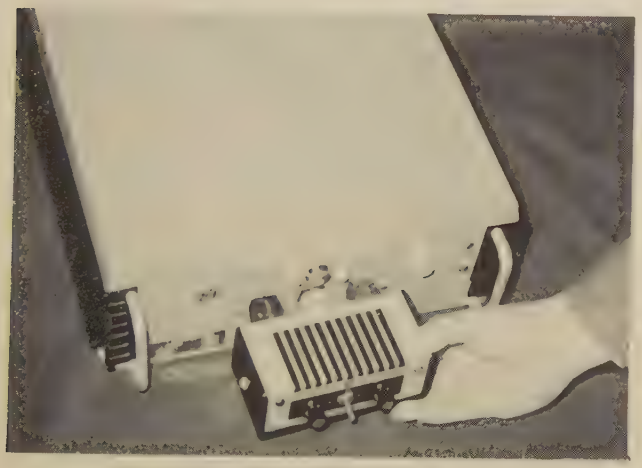
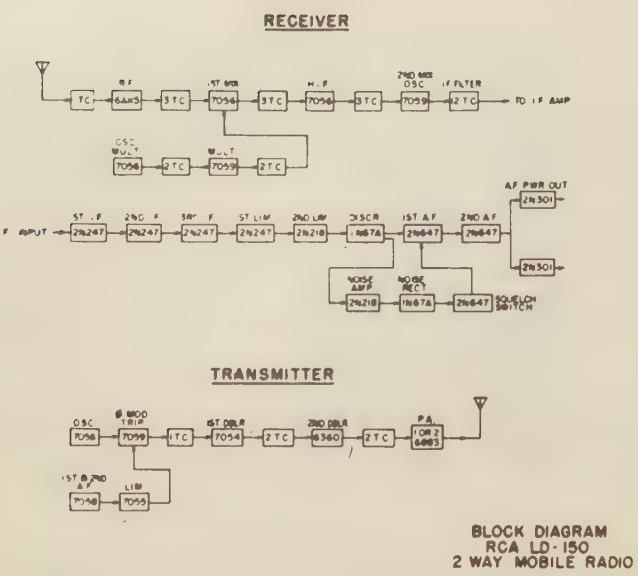


Fig. 3 - Conversion of Trunk Mount to "Under-Dash" Mount by Plug-in Control Box

Fig. 6 - Transistorized Portion of Receiver



BLOCK DIAGRAM
RCA LD-150
2 WAY MOBILE RADIO

Fig. 7

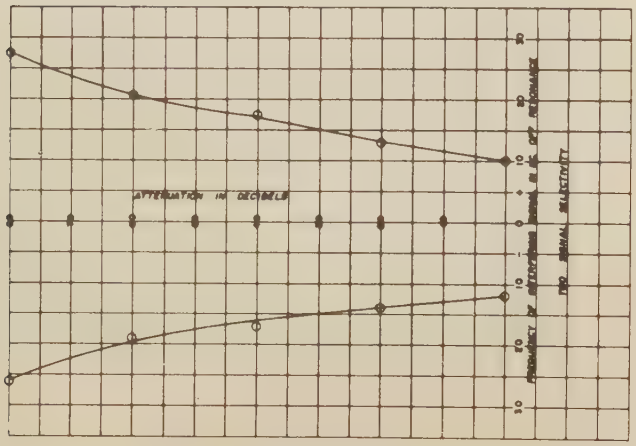


Fig. 8 (Left)

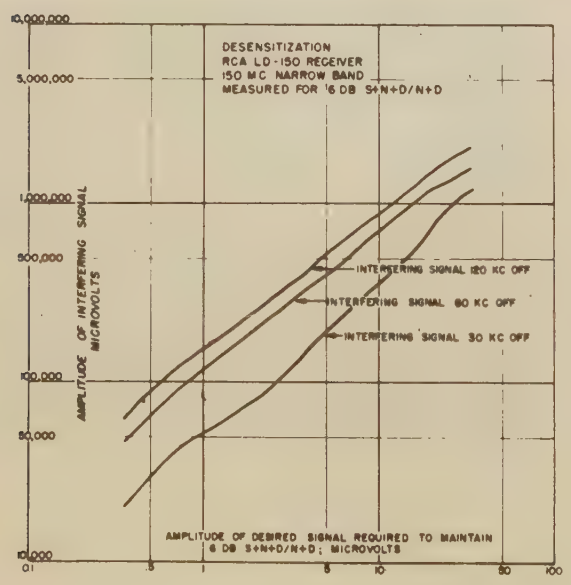

DESENSITIZATION
RCA LD-150 RECEIVER
150 MC NARROW BAND
MEASURED FOR 16 DB S+N+D/N+D
AMPLITUDE OF INTERFERING SIGNAL MICROVOLTS
INTERFERING SIGNAL 120 KC OFF
INTERFERING SIGNAL 60 KC OFF
INTERFERING SIGNAL 30 KC OFF
AMPLITUDE OF DESIRED SIGNAL REQUIRED TO MAINTAIN 6 DB S+N+D/N+D; MICROVOLTS

Fig. 9 (Left)

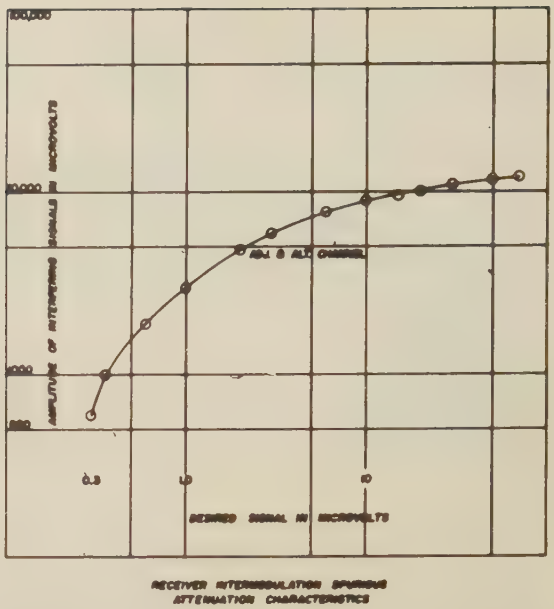

AMPLITUDE OF INTERFERING SIGNALS IN MICROVOLTS
ADJ & ALT CHANNEL
DESIRED SIGNAL IN MICROVOLTS
RECEIVER INTERMODULATION SPURIOUS ATTENUATION CHARACTERISTICS

Fig. 10

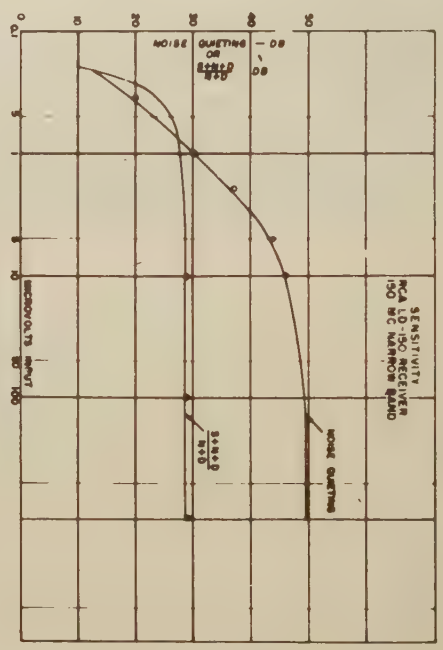


Fig. 11

COORDINATED BROADBAND MOBILE TELEPHONE SYSTEM

W. D. Lewis
Bell Telephone Laboratories, Incorporated
Murray Hill, New Jersey

Summary

The philosophy and broad outline of a coordinated broadband mobile telephone system are discussed. This system would be able to provide mobile telephone service of better quality than at present to many times the present number of customers. This could be made possible by means of an integrated switching and transmission design based upon a broad block of frequencies. All mobile units would be controlled over a data channel from a digital common control system at central.

The research, development, and production programs required to bring such a system into being would take a number of years, possibly as many as ten. Technical feasibility seems reasonably certain. To achieve economic feasibility might require a substantial effort. Realization of the system would have to be gradual and would require the cooperation of radio manufacturers, airplane and automobile manufacturers, and telephone companies.

Introduction

If future mobile telephone service is economical and of good quality, it is reasonable to expect a demand from many times as many customers as at present. Frequencies in the band around 800 megacycles now allocated to television would be usable for this purpose. Even if a substantial fraction of this band were assigned to mobile service, much more efficient use of frequency space than in present mobile systems would have to be made. In addition, a number of other requirements should be met. A grade of service comparable to that provided fixed customers should be given. The land-based equipment should be able to serve any mobile customer within range, regardless of his point of origin. It should be possible to establish a network of land-based central stations with overlapping coverage so that a customer in any well-populated geographical location can be served. The service should be largely automatic.

It is clear that a system capable of meeting all these requirements would be very different from any existing mobile system. Nevertheless, it is believed that such a system is technically feasible and in time can be proved economically feasible.

The first objective of the present paper is to discuss some of the technical characteristics of the high capacity mobile problem. The second objective is to suggest a system plan which can be used as a starting point for further discussion about the subject.

Any communication system which connects independently to a large number of remote points over a smaller number of channels must be both a switching and a transmission system. The switching, as in ordinary telephone switching, can be accomplished most effectively by modern digital techniques. This idea is applied to the present problem. A common control is provided at central. Central controls the mobile unit over a radio data link with a rate of 20 kilobits per second, or less.

The low bit rates required make possible the use of potentially very economical magnetic logical circuitry for most control functions of the mobile unit. Economy in the mobile unit is important since there is a large multiplying factor.

A system of this type requires a reliable data channel. This channel could be provided by modulating twin carriers which are then detected in a single circuit to provide a simple diversity system.

Ability to pack channels closely is limited by ability to achieve absolute frequency stability. It is assumed that this problem can be solved at central. The twin carriers of the data channel can be used in the mobile unit to develop a reference frequency for accurate selection of received and transmitted voice channels.

Finally some of the major technical and economic questions are mentioned.

General

Possible future customers for mobile telephone service may number in the millions on a nation-wide basis. If widespread service of good quality were given by means of radio techniques similar to those used today, far too much frequency space would have to be assigned to this use. The alternative is to provide a system for giving this service with a much more economical utilization of frequency space. This paper describes such a system in a preliminary fashion.

The sources of this system are many. Since World War II many organizations, including Bell Telephone Laboratories, have been concerned with mobile telephony. The ideas of coordinated utilization of a substantial block of frequencies and of continuous area coverage were being studied as much as ten years ago. Techniques for electronic common control and for data transmission, both relevant to the kind of system proposed in this paper, have been intensively investigated during recent years.

The frequency range between 450 and 890 megacycles is almost ideally suited for mobile telephone service. Propagation experiments[1] indicate that 450 megacycles is superior to 150 megacycles for mobile telephony, and that 900 megacycles is still reasonable. The higher frequencies extending into the microwave range are less desirable for a number of reasons.

(1) Shadowing and interference difficulties because of local obstructions, as in a city, are more severe;

(2) Achievement of adequate stability is more difficult because control to a smaller fraction of the absolute frequency is required;

(3) Transmission to, from or between moving stations is complicated by Doppler frequencies well into the audio range.

These difficulties as well as basic considerations of economy should lead us to the full exploitation of lower frequency ranges for mobile telephony before any attempt is made to use microwaves.

A number of means are known by which are more efficient exploitation of frequency space is possible than is now achieved in mobile telephony. These means if properly applied will also result in better service.

(1) Trunking: Present methods depend on assignment of a large number of mobile stations to a single, or at most to a few radio channels. This kind of assignment results in comparatively inefficient use of channels and a grade of service appreciably less attractive than wire telephone service, particularly from the speed of service standpoint. If the channels available to a mobile station are made numerous enough, perhaps 25 or 50, then statistically the channel usage is more efficient and the speed of service is superior.

(2) In a given geographical area, present mobile channels are spaced rather widely (at 150 mc - 20 kc and 450 mc - 50 kc). This spacing is necessary with the present system in order to permit adequate service without requiring frequency stability and selectivity which are too expensive. It also permits the use of F.M. with a larger deviation ratio, and thus lowers transmission requirements. The closer channel spacings will be possible if greater frequency stability can be cheaply achieved and if selectivity and transmission requirements can be reduced. Frequency stability is cheaper if a pilot frequency is continuously available at the mobile unit. Relaxed selectivity and transmission requirements are possible if channels are assigned on an integrated basis, taking advantage of geographical dispersion. These techniques are features of the system to be outlined.

(3) As in other radio services, geographical sharing of common frequency space is necessary for maximum frequency utilization. In a mobile system extreme anomalies in propagation make this more difficult. At these frequencies, particularly in built-up areas, multipath propagation can augment signals from remote transmitters and can cause signals from nearby transmitters to fade in a manner which makes discrimination by geographical separation much more difficult. Also, in any long-range mobile telephone system the "roamer" problem must be solved. When a moving station goes from one geographical area into another, the system should become aware of this and shift control of the call and frequency assignments accordingly. The system to be discussed can be extended to have these features.

Dollar economy is as important as frequency economy especially at the mobile station. It is clear that any system with the features suggested above will have to solve a number of complex control problems. The system to be described will reduce the control complexity required at the mobile station by placing as much of it as possible at central switching points. Common control equipment which would be required at central in any case for switching can be expanded to include the required control functions and the cost of this expansion can be shared among a very large number of mobile stations.

It should be possible in the proposed system to select equipment in the mobile unit to some extent of the type already used in broadcast radio or television. This would make it easier for radio manufacturers to produce the equipment and would reduce manufacturing costs.

Because of the integrated nature of the proposed mobile telephone system, particularly its inherent ability to select superior transmission channels at any time and provision of adequate data transmission, the system to be

---

[1] W. R. Young, "Comparison of Mobile Radio Transmission at 150, 450, 900 and 3700 Megacycles," B.S.T.J., November 1952, pp. 1068-1085.

proposed may be able to provide telephone service comparable in quality and flexibility to that provided now to fixed stations.

The mobile scheme proposed can ultimately be extended to include service to not only motor vehicles but to ships, trains, and to aircraft. The latter may require some special attention because of higher speeds and wider mobilities. Work on mobile telephony to vehicles may also lay the ground work for service to subscribers through individually carried pocket transceivers.

The general systems scheme outlined below could also have military applications. It is a scheme which would provide communication to highly mobile military forces including aircraft, tanks, personnel vehicles and individuals. It should provide far more flexibility, with less confusion, and probably at lower cost than any extrapolation of present military communication schemes.

The system described below is outlined in only the broadest fashion. Clearly the bulk of work, even on paper, still remains to be done. More detailed examination of the various possibilities may, and hopefully, will result in better configurations of the component units. Nevertheless we believe that the broad systems philosophy is a good one. This philosophy requires that a central control be responsible not only for telephone switching, but also for monitoring transmission and channel assignments to obtain maximum utilization of limited frequency space, and a high grade of service.

Traffic, Channel and Modulation Considerations

Preliminary consideration shows that the mobile problem is most difficult to solve in large metropolitan areas. Propagation experiments show that frequency sharing within a given metropolitan area might have some advantages in frequency economy. Such a plan will be discussed in the following paper by Cornell and Schulte.

Locations of large metropolitan areas in this country are sufficiently separate from each other so that each such area may use the bulk of available frequency space without serious risk of interference from other similar areas. In order to cover the country geographically, we might assign an initial band of about 24 megacycles to metropolitan service and eight megacycles to service in between. Additional bands of this size would be assigned as these become fully loaded.

Let us assume we must serve 18,000 mobile subscribers in a given metropolitan area with 24 megacycles. Of these perhaps 1/30, i.e., 600, will be busy with a mobile conversation in the mobile busy hour. This is 40 kc per conversation for transmit, receive, and control. How can we give a good grade of service with this tight frequency allotment? Present mobile telephone service at 450 megacycles uses 100 kc for the same functions. Such a lavish use of bandwidth is clearly forbidden.

Any simple application of pulse techniques appears out. There is far too little frequency space for P.C.M. P.A.M. requires a good degree of separation between successive pulses if cross-talk is to be avoided. In a city, successive pulses will be highly overlapped because of multiple path propagation.(2)

It would seem superficially simpler to use a single channel for both directions and to time-share on a push-to-talk basis or mechanized equivalent thereof. Such a scheme would, however, produce grave interference difficulties since it would require receivers which could protect against signals only a few kilocycles out of band and up as much as 150 db over the signals to be received. No practical RF filters or amplifiers are likely to be found which could sufficiently protect the first converter circuit of a receiver against overloading and interference from signals of this magnitude and proximity.

This kind of interference problem can be neatly solved by grouping land-to-mobile and mobile-to-land channels in sizable blocks separated by reasonable guard spaces. We also require some space for the transmission of control information.

Figure 1 shows a possible structure for one two-way circuit band. This could contain 600 metropolitan circuits and 200 rural circuits. Figure 2 shows a possible structure of one multichannel circuit block within this band.

Broad Features of the System

A block diagram of the system is shown in Figure 3. Transmission of voice must of course be through radio paths. These paths are transmitted and received in multiplex at the central station. In smaller areas or simpler situations, the central station may have at its disposal only a single radio transmit and receive location, but where there are many subscribers, it may be advantageous to have several such locations. Each central station will have available to it one or more blocks of frequency, depending on the number of mobile units which it must serve. Transmit and receive blocks will be kept separate.

The mobile unit will contain a voice transceiver capable of receiving and transmitting on any one of a number of, but not necessarily all, frequencies assigned to its central station.

---

(2) W. R. Young and L. Y. Lacy, "Echoes in Transmission at 450 Megacycles from Land-to-Car Radio Units," Proc. of the I.R.E., March 1950, pp. 255-258.

During a normal conversation the channel in each direction remains assigned. When the fixed subscriber is talking and the mobile subscriber is listening, the channel assigned to the fixed transmitter and mobile receiver will be activated, but not vice versa. For talking in the other direction, only the channel in the other direction will be activated.

In addition to the talking channel a highly reliable data channel will be sent out from the central station to all mobile units. Each mobile unit will contain a data circuit separated from the voice circuit after the intermediate frequency amplifier and tuned to this channel. This data circuit is sensitive at all times even when no call is in progress. Data from this receiver will be used to alert, integrate, assign frequencies to, and otherwise control the mobile unit in a manner to be discussed below. Responses of a mobile unit to interrogation are returned to central simultaneously over a suitable channel. The data channel will also contain within it in the form of carriers two pilot frequencies which can be used as frequency references for controlling absolute radio frequencies in the mobile transceiver.

Common Control

Some of the necessary functions of the common control unit are seen from the above skeleton description of the transceiver. The common control must assign suitable radio transmission paths in each direction. It must activate corresponding radio receiving and transmitting units in both the fixed and the mobile stations. There may be a single fixed station. In metropolitan areas in order to insure adequate coverage and good propagation there may be a multiplicity of fixed stations. The data channel serves as a link between the common control and all mobile units. Digital signals sent over the data link from the common control to the data receiver are used to control the state of the transceiver.

A call can be established from central to a mobile unit in the following manner. The characteristic code of the data receiver (possibly a full ten-digit number) is sent over the data channel to the data receiver. This readies the data receiver for further action. In addition, a code combination is sent to indicate that there is an incoming call for the mobile unit. This activates the bell or its equivalent in the mobile unit, alerting the subscriber. A code group assigning a mobile-to-fixed radio channel to the call is also sent to the mobile unit. This group permits selection and activation of a transmitting channel in the mobile unit simultaneously with the activation of the corresponding fixed receiving channel and also the connection of this channel to the correct fixed trunk at central. If the mobile subscriber indicates that he is ready by activating the switch hook or its equivalent at the mobile subset, then this fact is signaled to central over the assigned mobile-to-fixed channel. This mobile-to-fixed channel can be left assigned until the mobile called subscriber answers, or until the fixed calling subscriber decides there is to be no answer and hangs up.

When the common control receives information that the mobile called subscriber has answered, the fact that the conversation is established can be recorded in the common control and also, if desired, it can be signaled to automatic message accounting for charging purposes. A number of alternate systems designs for holding a call in progress can be worked out depending on the economics of common control complexity versus frequency utilization. In a large metropolitan area it will be desirable to achieve maximum frequency utilization through a high degree of common control intervention, whereas in a sparsely settled area, the reverse may be true. For reasons of compatibility the basic mobile unit should probably be the same in either case, though some economies might be possible in the simpler systems. The principal differences will be at central; but central will be asked to be more complex only where its cost can be divided among a very large number of mobile subscribers.

There are a number of ways to handle calls started by a mobile unit. One way could be as follows: the mobile unit would not be permitted to transmit until it had been assigned a transmitting frequency; yet it would be wasteful to assign it a transmitting frequency at all times regardless of need. The common control by means of the data channel would interrogate the mobile units for which calls are not in progress. A mobile subscriber wishing to initiate a call would activate his switch hook or equivalent. With the mobile unit in this condition, the interrogating signal would cause the mobile transceiver to transmit a demand for service signal to central. Central could now proceed to identify the mobile unit desiring service. Once central knew which mobile unit desired to be active, the call could proceed more or less as before.

The common control unit has access, directly or indirectly, to all information affecting the integrated mobile telephone system under its general charge and can send out control signals to all fixed and mobile equipment. It must perform all of the functions of the common control box. In addition it can assume in a determined controlled switching system responsibility for supervising the condition of radio transmission paths and assigning them as needed to provide optimum service. In conditions of light traffic it will be able to select channels with greater frequency spacing to reduce interference. Likewise, it can select preferred fixed transmission and receiving locations where a multiplicity of these are available.

Technical and Economic Problems

Clearly, what I have outlined above is a broad system plan rather than a detailed design. Furthermore it is in the nature of a long range objective, rather than a description of something that is in existence today or can be in existence in the near future.

No insoluble technical problems need be solved to bring such a system into being. The complexities of the common control system are no greater than those of a special purpose digital computer or of experimental electronic switching systems already in existence. Receiving, transmitting, and frequency stabilizing equipment capable of meeting the technical requirements exists in a number of forms. The control unit of the mobile set can be made of transistor or even of magnetic logic circuits for which technical feasibility has been proved many times over. Perhaps the most difficult technical problems are those of reliable transmission, especially of data; but even here, methods of diversity and redundancy which have been amply demonstrated can be brought effectively to bear.

The most difficult problems are those of economics. How can we perform the complex functions required at a low enough price to be attractive? We are some distance from there now. This is the goal to which our most intensive research and development must be directed. To help us, there are two strong price reducing forces at work:

(a) Expansion of mobile service will mean mass production.

(b) Solid state techniques are developing and becoming cheaper at a rapid pace. A paper in these Proceedings by L. G. Schimpf tells you about a solid state receiver in the 850-900 megacycle frequency region.

If this kind of mobile telephone service becomes as widespread as we believe it can and should, it will provide considerable business for the telephone companies as well as to many radio and automobile dealers and manufacturers throughout the country. Mr. Huntley, Chief Engineer of the American Telephone and Telegraph Company, has aptly expressed the Bell System's attitude regarding the division of responsibility for furnishing the base station and mobile equipment in the following words:

> "The principal part of the telephone companies in providing mobile service would be to do the over-all systems planning and to furnish the base station facilities and connection to the telephone network.
>
> "The mobile radio units would be designed and manufactured by the radio industry and usually owned by the vehicle owner. It seems likely the radio set would be regarded as an automotive accessory and furnished on a basis similar to other accessories."

Conclusion

The philosophy and broad outline of a coordinated mobile telephone system have been discussed. This system would be able to provide mobile telephone service of better quality than at present to many times the present number of customers.

Such a system could not come all at once, but would require a research, development and production program lasting perhaps as much as ten years. It would have to come as the result of an orderly and not too abrupt transition from what we have now.

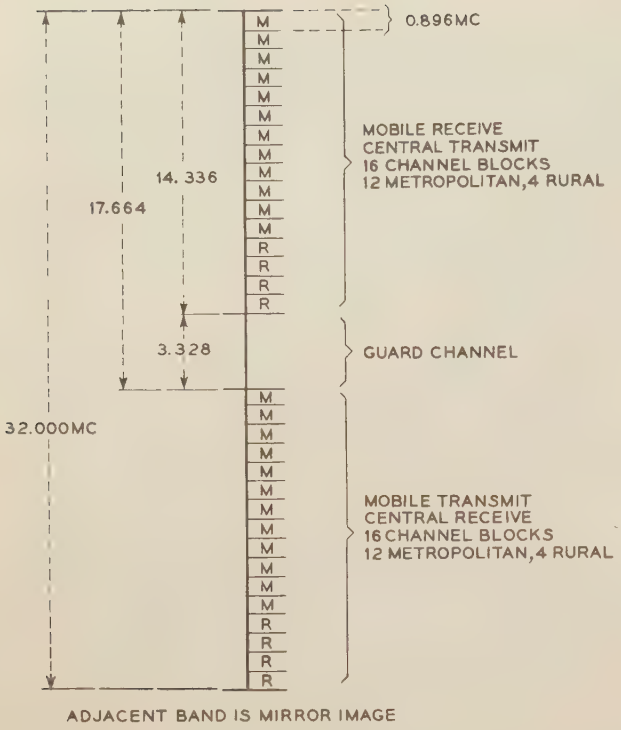


FIG. 1
FREQUENCY DIAGRAM
STRUCTURE OF ONE MOBILE BAND
IN 450-890 MC REGION

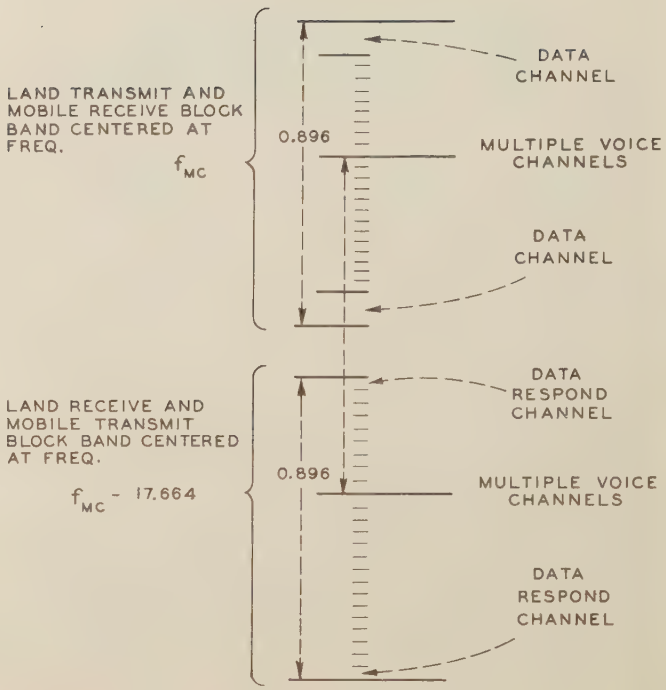


FIG. 2
POSSIBLE STRUCTURE OF MOBILE
CHANNEL BLOCK

HIGH CAPACITY MOBILE TELEPHONE SYSTEM
BLOCK DIAGRAM

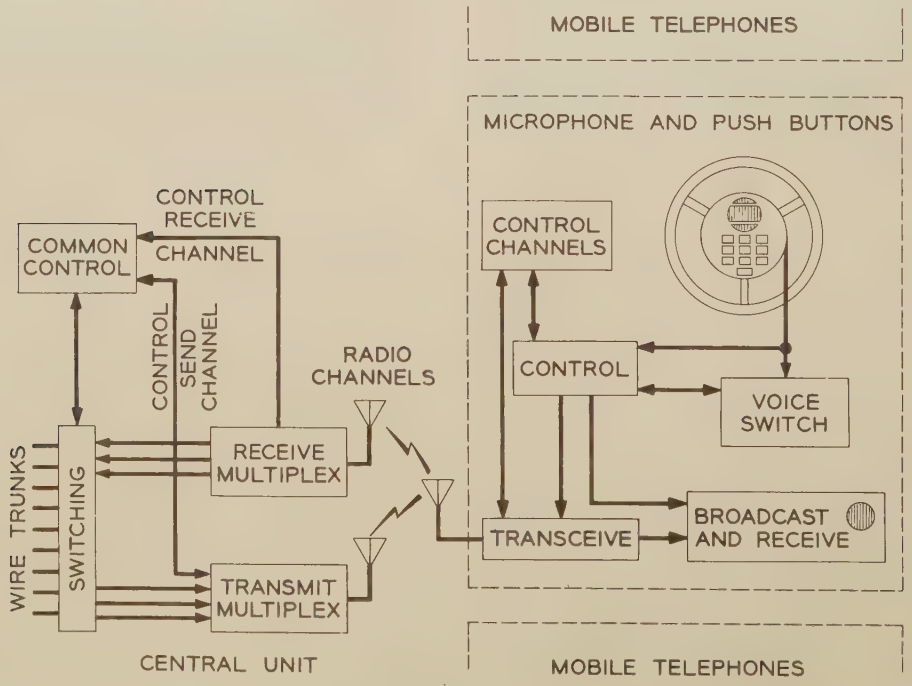


FIG. 3

MULTI-AREA MOBILE TELEPHONE SYSTEM

by

H. J. Schulte, Jr. and W. A. Cornell
Bell Telephone Laboratories, Incorporated
Murray Hill, New Jersey

Summary

A research study directed toward a broadband, coordinated, mobile telephone system is reported. In particular, a system using many fixed transmitter-receiver installations is proposed. Such an arrangement has three principal advantages: (1) economy in spectrum usage, (2) relaxation of interference requirements in the fixed and mobile equipment, and (3) reduction of transmitter power. In order to gain these advantages, it is necessary to provide an interconnection facility for these many remote stations. Considerable use must also be made of modern high speed switching and common control techniques.

Introduction

A broadband mobile telephone system capable of providing greatly expanded service by means of switching techniques and high speed common control has been discussed by Lewis in the previous paper.(1) Using this as a point of departure, we shall discuss further the possibility of using both space and frequency separation to handle large numbers of mobile subscribers which are expected in the near future.

The study upon which this talk is based is a continuing one. These remarks must be regarded as purely preliminary and exploratory in nature. In attempting to modify the system proposed by Lewis in such a fashion as to reduce spectrum usage and ease intermodulation requirements, we raise perhaps as many questions as we answer. In particular, it becomes clear that further study of the propagation problem is needed. It is believed, however, that the advantages of such a multi-area type of system make further study worthwhile.

In conventional mobile telephone service, simultaneous conversations in a single metropolitan area are handled by frequency division on separate radio channels. These channels are then re-used in different geographic areas where the space separation is great enough to prevent co-channel interference. Ring, Bullington, and others at Bell Telephone Laboratories have previously discussed the possibility of exploiting the space separation idea.(2) For a given traffic density (simultaneous calls per square mile) the number of channels, and hence the bandwidth required, ideally is proportional to the square of the coverage radius of a single transmitter. In principle then, one can achieve great spectrum economy merely by use of many fixed stations, separated by a fraction of a mile rather than by tens of miles as at present. Obviously, the power radiated by the fixed and mobile stations can also be reduced ten-fold or more.

However, there is one major disadvantage: this saving in bandwidth and radiated power must be paid for in increased system complexity. Hence lest the spectrum needed for mobile services be underestimated, we must point out that this complexity also ultimately raises the cost to the user and therefore as in most things, a compromise must be struck.

The complexity comes about in two major ways. First, the voice circuits themselves must be extended from the control center to each of the remote multi-channel stations. Secondly, there must also be coordination between the remote stations since there will be many mobiles which cross sub-area boundaries, even during conversations. There must also be coordination between control centrals to minimize interference and to handle the roamers. This coordination is feasible with present day switching techniques and does not require an excessive bandwidth for the supervisory operations.

To illustrate the proposal, the general layout of a metropolitan system is shown in slide 1. Three co-channel areas are outlined, within any one of which no two conversations use the same channel. Each co-channel area is sub-divided into many sub-areas. In each of these a satellite transmitter receiver is located. The network which connects them to their control central is shown for one co-channel area. This interconnecting network includes the voice circuits and such supervisory circuits as are needed to coordinate the fixed remote stations and through them, the mobile units as well.

There is another important advantage of such a space division system which has not previously been recognized. This derives from the ability to program the control central to select frequencies for each sub-area which are mutually non-interfering. Since we have to separate sub-areas which are to use the same channels by a distance of at least six coverage radii, we use the remaining frequencies, those which would produce interference, in the surrounding sub-areas. This is shown on slide 2. From a block of 500 equally spaced channels, we pick 20 or so which produce no bothersome third order intermodulation products. These are assigned to sub-area a. The next set of 20 chosen from the remaining 480 channels

are used in sub-area b, and so on. Thus a correlation of the required co-channel separation and the numerology of picking non-interfering channels can produce a highly efficient system.

As a direct result of the spectrum economy accruing from the reduced coverage radius the system designer is more or less free to choose modulation methods, redundancy in supervisory transmission, etc., with emphasis upon dollar economy and reliability. He no longer is forced to use the minimum bandwidth modulation method if some other system is more attractive economically and operationally.

Another important advantage of the system is its ability to grow as traffic demands change over the years. It is relatively simple to split one sub-area into two or combine two into one by changing the interconnecting network and moving transmitter-receiver sites. In the remainder of this paper, we shall discuss the effects of this type of system upon the various components and circuits.

Discussion

A. Frequency Plan and Selection of Sub-Area Size

The problem of intermodulation interference has been recognized since the early days of mobile radio. It still will remain as a rather serious consideration, especially in broadband systems. The task of transmitting 100 channels at 20 watts apiece without undue intermodulation is formidable indeed. There are potentially 3,800 third order interfering products superimposed on each channel. One solution is an ultra-linear transmitter amplifier. There are other possibilities of comparable or even greater cost which we will not discuss. In addition, there are difficulties in both the base and mobile receiver. The solution to be considered here involves not using the entire band. Instead suppose we use only those frequencies which produce intermodulation products which fall into unused channels. This is rather easily done; the selection scheme has been discussed before by Babcock.(3) To avoid third order products, for example, one merely uses those channels out of the original 100 which have no common intervals of frequency separation as shown on slide 3. The only difficulty is that this scheme only yields 11 usable frequencies.

Obviously we must use the remaining channels elsewhere and the obvious place to use them is in the sub-areas surrounding the original sub-area. If we choose the number of sub-areas to be approximately 100 divided by 11, we might get almost complete utilization of the original 100 channels, at the rate of about 11 channels in each of nine sub-areas. The area including these nine sub-areas is the co-channel area. Hopefully, these same 100 channels can then be re-used outside this region.

Thus far the number of channels has been quite arbitrarily set at 100, perhaps because that is the number which Lewis mentioned. There are several conditions which we can impose upon the solution, however, which help to take some of the arbitrariness out of the problem. First of all, it has been assumed that the co-channel area contain enough equal size sub-areas to permit re-use of frequencies outside of the co-channel area. Six is perhaps a minimum ratio of co-channel separation to coverage radius for FM systems; somewhat more is required for AM. This is discussed further in the next section. Purely from geometrical considerations, this requirement suggests a minimum of about 16 sub-areas per co-channel area. Twenty is selected as a practical minimum to permit some size variation in the smaller areas.

Secondly, to be able to choose about the same number of non-interfering channels for each sub-area, it is necessary to make the number of sub-areas equal to the total number of channels divided by the number of interference-free channels which can be selected from that total number. Slide 4 shows the number of interference free channels, p, which can be selected from the total number of channels, n, for elimination of third order products. The steeper line is drawn for a system having 20 sub-areas, that is $n/p = 20$, which is about the minimum. We are thus led to choose a system having 20 sub-areas per co-channel area and about 500 channels total.

The third condition which fixes the scale factor of the system is the traffic density, expressed as the maximum number of calls we expect the system to handle per square mile. We see again from slide 4 that the 500 channel system provides a maximum number of calls, p, of about 25 per sub-area. Hence if we expect a maximum calling rate of 50 calls per square mile we must make the sub-areas about one-half square mile in area.

One further remark before leaving the question of sub-area size selection. If one attempts by enumeration the assignment of channels in such a multi-area system, it soon appears that it is not possible to select an equal number of non-interfering channels in each sub-area. The actual number available to the last sub-area assigned is about one half the number possible in the first. This is no particular disadvantage, since the peak calling rate will probably not appear simultaneously in all 20 sub-areas. Therefore, with the flexibility of fast switching and common control the system will assign the largest group of channels to the sub-area with the most traffic at any given time. It is also possible in emergency situations or times of very heavy traffic to permit assignments to be made beyond the above non-interfering scheme. This will of course degrade the quality of transmission, and must not be permitted to make any gross inroads upon intelligibility.

B. Antenna Design and Siting

In conventional mobile systems the base station antenna is usually located on a tall building or on a hill overlooking the city. Path loss variation over the desired coverage area is reduced by a limited amount of vertical directivity which beams more of the energy toward the horizon than straight down. Slide 5 illustrates the general philosophy, showing the antenna radiation pattern and the resulting path loss behavior with distance. Note the slow variation (less than 2 db/octave) out to one mile radius, followed by a more rapid fall off as first free space and then earth diffraction conditions prevail.

A co-channel to coverage radius ratio of three or so was suggested by Bullington[2] for systems whose coverage radii were about 20 miles. This is seen to be reasonable from slide 5. However, explicit in earlier parts of this talk was the need to shrink the distance scale by at least a factor of ten. This requires a different approach to antenna design and siting.

A very low to the ground base station antenna would provide steeper path loss variation. A slope of 12 db/octave is possible with an antenna which is low enough to obtain plane earth conditions. A further possibility is a higher, directive antenna with depressed lobes to beam most of the energy down onto the desired area. Other possibilities include locating the antennas in the "valleys" between the "hills" rather than on the peaks as is now the rule.

As was indicated in the introduction, there is need for considerable field work to establish the feasibility of such path loss shaping in a real world. Further, it is clear that the installation of such a multi-area system involves a systems engineering endeavor of a high order.

C. Effects on Mobile Transceiver Design

In order to achieve trunking advantage, it is necessary that the mobile set be able to receive and transmit on several of the channels used by each of the sub-areas. This immediately leads to the requirement that the mobile transceiver be capable of operating on a large number of frequencies under the control of a signal from the control central. If we assume that the mobile set must be able to use a minimum of five channels from each sub-area, it must be capable of being switched to one of 100 channels for a system having 20 sub-areas. This may not be too difficult a requirement to meet since once means are provided for remote channel selection, the number selected may not be too important.

Linearity requirements in the mobile receiver are minimized since it is not possible for a mobile subscriber to be listening to one transmitter and receive strong signals from a second multi-channel transmitter which could form interfering intermodulation products in the front end of his receiver. In such cases the subscriber would be switched to a channel from the transmitter having the highest signal level at his location. By avoiding the use of adjacent channels in any one base station, the selectivity requirements of the mobile receiver are also relaxed.

There appears to be no reason why mobile receiver sensitivity should be better than that used in present mobile systems (about one microvolt for 20 db of noise quieting). On the contrary, unless some sort of suppression is added, impulse noise may be limiting, especially in the data signals. Sensitivity figures of five microvolts or so may therefore be more realistic. This, of course, is helped by the use of small coverage areas from the base transmitter which permit laying down a stronger signal without the use of extremely large transmitter powers.

Radiated power per channel from the mobile transmitters should be about the same as from the fixed stations since only one fixed receiver will be provided for each sub-area transmitter. The maximum power required is determined by the size of the largest sub-area in the system and by shadow loss and multipath fading. Since the largest sub-areas will occur in suburban or rural areas where propagation may be better than in a city, there may be some equalizing factor. In any event, it is desirable that the mobile power radiated be controlled from the control central for reasons discussed in the next section.

D. Effects on Base Station Design

Whereas the mobile receiver need receive only one talking channel at a time, the fixed receiver must receive all channels in use within its coverage area simultaneously. This normally tends to increase requirements on linearity in stages ahead of the channel separation filters. These requirements are, of course, lessened by the selection of non-interfering channels. Furthermore, when, because of heavy telephone traffic, channels are used which can cause intermodulation, an advantage can be obtained by using a minimum signal level at the fixed receiver. This can be accomplished by controlling the power output of the mobile unit in response to supervisory signals over the data channel. In addition to reduction of the amplitude of

products generated in the fixed receiver, this will tend to reduce blocking by nearby mobile transmitters by insuring that the received signal does not exceed a predetermined level. Because of its more favorable location with respect to noise sources the sensitivity of the base receiver can probably be made greater than that of the mobile receiver.

The base transmitter must be capable of transmitting on any p of n channels simultaneously, where n may be a large number; thus the use of separate transmitters does not appear attractive. This leads to the requirement for a transmitter with a degree of linearity sufficient to operate on a few tens of channels simultaneously without the generation of unwanted products. If channels are selected in accordance with the plan previously described, in-band products falling on working channels are eliminated. However, there still exist out-of-band products and in-band products falling on channels used by an adjacent transmitter. If we make full use of channel selection, a mobile will always be receiving from the base station which provides it with the largest signal. Therefore, intermodulation products from a second transmitter will arrive at a lower level.

E. Data Link Considerations

It is planned that all mobiles in one co-channel area receive the same data signals from the control central. It is a requirement that when a mobile goes from one co-channel area to another he then is able to receive data signals from the new co-channel area.

Two types of data messages are required: general address messages which must be sent to all sub-areas within a co-channel area, and messages addressed to a specific mobile which could be sent only to the sub-area in which the car is located. One method would be to operate the data transmitters in adjacent sub-areas on different but closely spaced frequencies with the same data modulation on each. The mobile receiver must then be able to receive any or all of these frequencies. A similar scheme for handling messages has been proposed by Brinkley.[4] This method, of course, will not work at the boundaries between co-channel areas where the messages are different. One method of handling both problems would be to go to a time division arrangement on the data channels, sequentially signalling the various sub-areas in each co-channel area.

Concluding Remarks

The aim in this presentation has been to provide a somewhat simplified picture of some system ideas which are in themselves and individually not particularly new. However, the use of small coverage radii from low power transmitters in combination with a coordinated, non-interfering frequency pattern is only now becoming feasible through the recent advent of fast switching and common control techniques. It should also be repeated that these techniques and the system as a whole will not be without cost to the user, hence a compromise between spectrum economy and dollar economy is called for as usual.

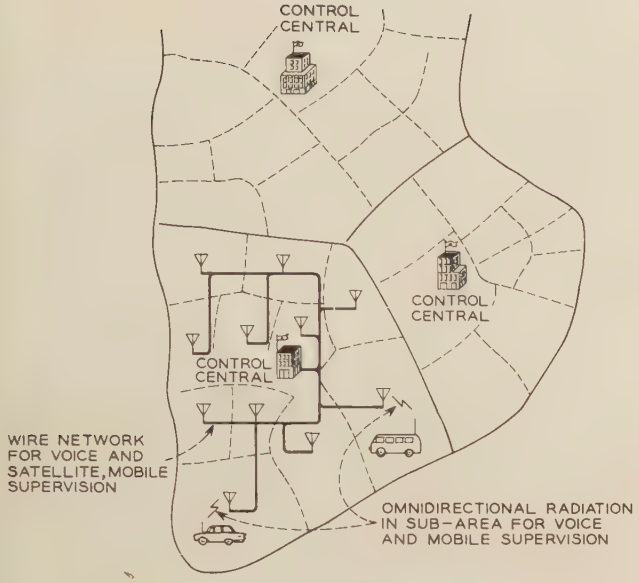
References

1. Coordinated Broadband Mobile Telephone System, W. D. Lewis (previous paper).

2. Frequency Economy in Mobile Radio Bands, K. Bullington, B.S.T.J., vol 32, page 42 (January 1953).

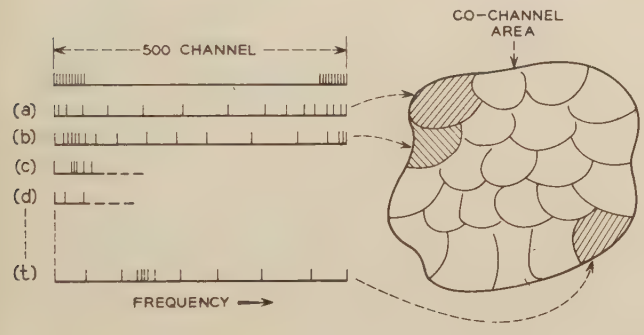
3. Intermodulation Interference in Radio Systems, W. C. Babcock, B.S.T.J., vol 32, page 63 (January 1953).

4. A Method of Increasing the Range of VHF Communication Systems by Multi-Carrier Amplitude Modulations, J. R. Brinkley, J.I.E.E. vol 93, part III, page 159 (May 1946).


CONTROL CENTRAL
CONTROL CENTRAL
CONTROL CENTRAL
WIRE NETWORK FOR VOICE AND SATELLITE, MOBILE SUPERVISION
OMNIDIRECTIONAL RADIATION IN SUB-AREA FOR VOICE AND MOBILE SUPERVISION

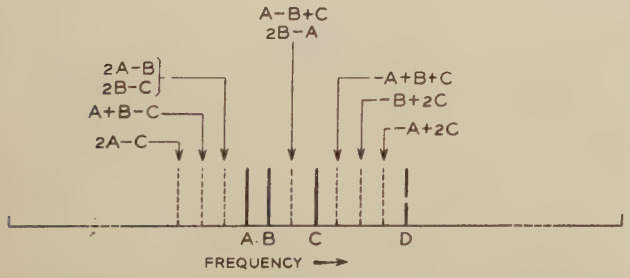
Slide 1

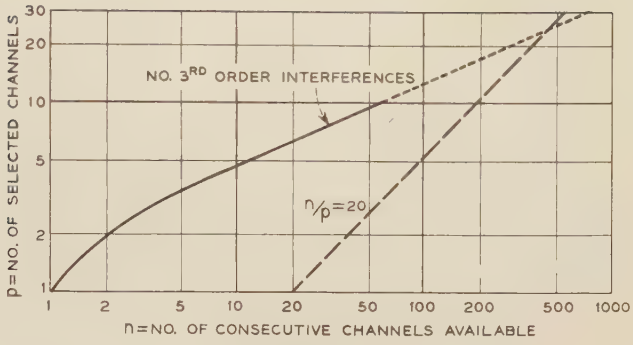
Geographic layout of multi-area system showing control centrals, sub-areas, and interconnection network.



Slide 2

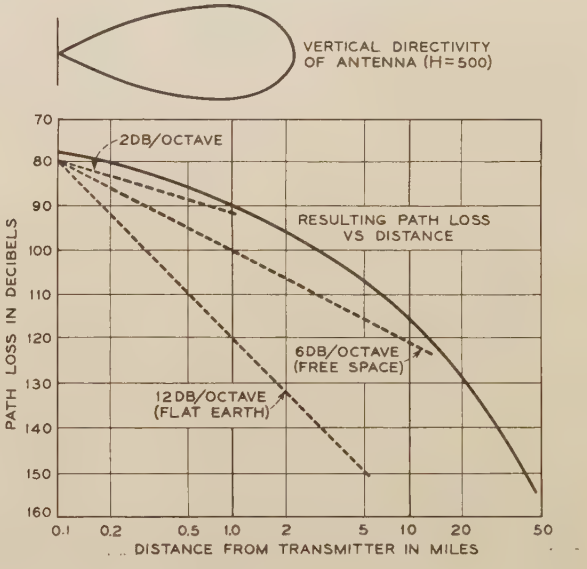
Frequency assignments in a co-channel area. No mutually interfering channels are used in any single sub-area.





Slide 4

Number of non-interfering channels available from a given set of equally spaced channels. A near optimum system is obtained at the intersection of the two curves. The line for $n/p = 20$ relates to a system having 20 sub-areas per co-channel area. (After W.C. Babcock, B.S.T.J. 32, pp. 63, January 1953.)



Slide 5

Typical conventional antenna engineering. Vertical directivity beamed at the horizon and a high location are assumed. The resulting path loss vs distance curve is shown below, neglecting fluctuations due to shadow and multipath fading effects.

Slide 3 (Left)

Troublesome third order interference products generated by channels A, B and C. A fourth channel would be placed at frequency D.

PRECISION CARRIER FREQUENCY CONTROL AND MODULATION PHASE EQUALIZATION OF BASE TRANSMITTERS IN A MOBILE RADIO SYSTEM

D. S. Dewire, Supervising Engineer - Radio and TV
E. A. Steere, Supervising Engineer - Transmission
New York Telephone Company
Albany, New York

Summary

Distortion-free reception by mobile units of base transmissions is essential in multiple base station mobile radio systems, where adjacent stations are operated simultaneously and modulated with the same intelligence. Phasing and heterodyne distortion factors in overlap coverage areas must be minimized. After many months of laboratory and field testing, a satisfactory solution has been developed and applied to a system using wire lines for base station and control point interconnection. This method provides precision control of carrier frequencies, and phase equalization of modulation by building out the wire lines so that transit times are equal. After two years of operation this method has proven to assure satisfactory distortion-free reception.

INTRODUCTION

Multiple Base Station Radio Systems

Many mobile radio systems require the use of two or more adjacent base stations which, in order to obtain continuous coverage, must be operated simultaneously and transmit the same modulation. It is often impossible to determine the location of a mobile unit in respect to a particular base station. It is also usually impractical to operate base stations individually with assurance that the desired mobile unit is within range of that station.

There are two general types of multiple base station systems: a ribbon-type system, such as that required on a continuous express highway; and an area-type system, such as that required for police operations throughout a State. On ribbon systems following a relatively straight path across relatively level terrain, there will normally be overlap operation between two adjacent base stations. To cover the continuous path between these base stations at a signal level adequate for satisfactory operation, these overlap areas may be of appreciable length. Where the ribbon system must conform to rough terrain, the overlap areas are usually not well defined, but will depend on reflection effects from terrain irregularities, which may vary with atmospheric changes. Also, changes in direction of the highway may create areas which will experience overlap from more than two base stations.

The Heterodyne Problem

It has been found that reception by a mobile unit of signals from two or more base transmitters, transmitting on the same nominal RF frequency and differing in strength up to about 30 db, results in a heterodyne or beat frequency equal to the difference in actual RF frequencies of any two base transmitters. If these base transmitter frequencies are within approximately 3,000 cycles of each other, the heterodyne frequency is accepted by the mobile receiver and appears as an audible tone in the loudspeaker. This tone can be annoying in varying degrees to various people, and in many cases can be sufficiently strong to effectively mask an incoming message.

A method has now been found and applied which has, for all practical purposes, eliminated masking and annoyance from heterodyne. This method will be described in detail in the second portion of this paper.

The Phasing Problem

With FM modulation, where the signal strength from one base transmitter exceeds the signal strength from any other transmitters by about 6 db, the stronger signal will "capture" the receiver of a mobile unit. A clear signal from this stronger transmitter then obtains, and no difficulty in intelligibility is experienced. Where signal levels from two or more transmitters vary by less than the "capture" amount, both signals are received and distortion is usually present due to variations in phase of modulation of the transmitters. Many methods for minimizing such phase differences have been discussed and tried. The final portion of this paper will cover a method which has been proven in service to adequately minimize such phase distortion, where modulation supply to base transmitters is carried by individual wire lines, so that satisfactory reception by a mobile unit is assured.

Earlier Methods

A number of methods have been attempted to minimize these effects. One of the most effective in minimizing both heterodyne and phase distortion has been the use of dual antennas on the mobile unit, arranged in line with the car and spaced about 1/8 wave length apart. One antenna is active

while the other is passive and acts as a director. By means of a switch in the car, the antennas can be reversed in operation. Thus an appreciable directional affect is achieved so that interaction areas can be quite successfully circumvented by intelligent operation of the switch. However, there are limits to the use of such an arrangement. First, intelligent cooperation of the radio operator in the car must be assured. Second, such an arrangement can be used successfully only on ribbon systems following a straight path over level terrain. It is not satisfactory where curves exist so that the unit is often broadside to the propagation path, where areas must be covered by base stations which are not in a direct line with the route of travel, or where irregular terrain results in multiple reflections from various directions.

The extent of interaction areas can be reduced by close spacing of base stations so that signal strengths are kept at a very high level throughout the coverage area. This, however, is seldom economically feasible, and does not, in any event, completely minimize the distortion effects.

Other attempts at minimizing heterodyne have involved frequency adjustments, at adjacent base transmitters, wide enough apart (while still within the required frequency tolerance) so that any resultant heterodyne would be above the audio pass band of the mobile receiver. Without precise frequency control, normal frequency drift will prevent any assurance that the heterodyne can be kept outside the audio pass band. This is even more difficult with the decreased frequency tolerance under narrow-band operation. Further, heterodyne reception within the phase interaction zone may result in some additional speech distortion.

PRECISION CARRIER FREQUENCY CONTROL

Requirements

As mentioned above, experience on working systems shows that audible heterodyne is distinguishable in a mobile unit where transmissions from two or more base transmitters are received at strengths differing by up to 30 db. Experiments conducted by the Bell Laboratories have shown that within these limits, a heterodyne frequency between 150 cycles and the upper limit of the mobile receiver audio pass band is plainly audible in a loudspeaker or handset receiver. Heterodyne frequencies under 150 cycles are effectively attenuated below the annoyance or masking level. Heterodyne frequencies under 50 cycles, however, may result in periodic squelch operation in the receiver due to the slow rate of alternate signal reinforcement and cancellation. This squelch operation may cause "shot" noise effects which may be even more annoying than an audible tone.

Optimum performance in minimizing heterodyne therefore required that the transmitter RF frequencies of the contributing base transmitters be held to within 50 to 150 cycles of the assigned frequency. It was realized that it might be difficult to attain and maintain stabilization of frequency within these limits. Since it would be impractical to provide continuous manual adjustment, it was further required that the frequency adjustment hold for at least a month under wide ambient temperature variations.

The problems involved in obtaining such stabilized frequency control and maintaining it in accordance with requirements were many and difficult.

Crystals

Precise grinding of crystals was necessary beyond the precision normally expected. Many different crystals were tried. A crystal that met the requirements was finally obtained from Reeves-Hoffman in Carlisle, Pa. 85° C crystals were first used on a trial basis, but were later replaced with 75° C crystals.

Ovens

In order to hold the crystals to the required frequency limits, it was, of course, necessary to place each crystal in an oven, thermostatically controlled to keep the crystal temperature as near constant as possible. A number of ovens were tried with different thermostat controls. After many months of laboratory testing, an oven manufactured by the Hunt people was tried in service. This oven operated on the normal 6V heater supply. The original thermostat used with the oven did not operate satisfactorily under field trials, since the contacts had a tendency to stick either in the closed or open position. Further there was no external indication as to whether the thermostat was functioning correctly, until heterodyne appeared and was reported. Again laboratory tests were instituted and it was finally found that an Edison thermostat, operating on 117V supply and arranged to retain a substantially constant temperature of 75° C, appeared to be thoroughly reliable. It was, therefore, necessary to modify the Hunt oven to use an Edison thermostat and to provide a 117V supply. The new ovens (See Figure 1) were then placed in service and have operated satisfactorily for nearly two years.

Where base transmitters are housed in heated quarters, so that room temperature is fairly constant the year round, the ovens with the Edison thermostat and 75° C crystal will hold an adjustment within the required limits for months at a time. In unheated metal huts with normal summer to winter temperature variations, crystal frequencies will hold within required limits for

at least a month without readjustment.

Temporary frequency excursions beyond required limits are experienced when the equipment cabinet door is opened or closed. However, thermostatic action returns the frequency to normal within a very few minutes. These excursions can be minimized by the provision of heat insulating enclosures around the oven, but with normally unattended equipment the excursions are so infrequent and of such short duration that this treatment is not justified.

The oven, with its crystal, and the oscillator are mounted on a small platter (See Figure 2), which in turn is mounted in the space on the transmitter chassis normally reserved for a second frequency oscillator (See Figure 3). The standard crystal is left in place in the usual position and can be readily placed in service temporarily in case it is necessary to replace a stable oscillator in which trouble has developed. In order to quickly tell whether an oven is functioning properly, a neon light is provided on the stable oscillator platter, which glows whenever the thermostat closes to apply power to the oven heater.

Power Supplies

The 117V power supply for the oven and a regulated 6V supply for the oscillator heater filament are obtained from a separate small platter (See Figure 4) with a power transformer, rectifier, ballast tube, and regulator tube. This power supply platter may be mounted as convenient (See Figure 5) and connected to the stable oscillator platter (See Figure 6).

Complete units with stable oscillator and power supply are now available from the General Electric Company.

Test Equipment

Precision Frequency Meter - In order to obtain the critical adjustment required of the stable oscillator, test equipment used must be of even greater precision than the oscillator itself. Since similar precision adjustments are required on some air-ground applications, our Bell Laboratories people had already requested the Budelman Company to undertake design of a frequency meter for this purpose. The Budelman 40A Frequency Meter was the result of this work (See Figures 7 and 8). This meter is designed to measure two frequencies within the range of 150 to 175 mc with an accuracy of .000035%, or about 45 cycles at the frequency in use on the system where our stable oscillators are in service.

A signal is introduced into the meter from the transmitter and is compared with the nearest harmonic of a 2.5 mc crystal oscillator. This harmonic must be more than 1 mc and less than 2.5 mc away from the frequency to which the transmitter is to be adjusted. The difference frequency from this comparison is then compared with a second crystal oscillator, which will produce a difference frequency in the neighborhood of 1,000 cycles. A 1,000 cycle frequency applied to the deviation meter gives a zero reading on the meter. Any variation from 1,000 cycles can be read directly on the meter, which is calibrated from -250 to +250 cycles. The stable oscillator can then be adjusted so that the incoming signal from the transmitter will give a center reading on the meter.

WWV Receiver - Since no meter can be expected to give absolute accuracy under all conditions, it is necessary, first, that the meter be allowed a warm-up period sufficient to stabilize its operation and, second, that it be calibrated against a known source of recognized reliably exact frequency. Such a source is available in the carrier frequencies of WWV transmissions. Therefore, a WWV receiver is a required test instrument for calibrating the 40A meter. A simple antenna for receiving WWV is provided at each base station. A Specific Products WWV receiver is used (See Figures 9 and 10). This receiver is equipped with a dial which allows selection of any WWV frequency, and that on which the best and most fade-free reception is obtained is the frequency which is used. This WWV frequency is then compared with the output of the 2.5 mc oscillator in the 40A meter, and the oscillator adjusted accordingly.

Transmission Line Coupler - In order to secure enough signal from the output of the base transmitter for use with the 40A meter, and still not interfere more than necessary with normal operation of the transmitter, a special coupler (See Figures 11 and 12) is used in the transmission line between the transmitter and the antenna. This coupler consists of a special T connector with the pin in the male branch cut off at the level of the polethylene insulation. The portion of the pin which was cut off is then inserted in one end of a double female connector, the other end of which is connected to a coaxial cord to the 40A meter. When the end of the female connector with the pin inserted is connected to the T connector branch with the pin cut off, a capacitive coupling is effected which can be varied to obtain the necessary amount of signal for operation of the 40A meter.

Adjustment Procedure

The adjustment of stable oscillators requires use of the test equipment listed above; i.e., WWV receiver, 40A frequency meter, and RF coupling unit, and an ST13A (or equivalent) frequency meter (See Figures 13 and 14). Power is

applied to the frequency meters and WWV receiver, and at least 15 minutes is allowed as a warm-up period. Meanwhile the RF coupling unit is inserted in the transmission line between the transmitter and its antenna, and connected to the ST13A.

After warm-up, the transmitter is energized while a preliminary rough adjustment of transmitter frequency is made with the ST13A to assure later reference to the proper oscillation mode in the 40A meter. The ST13A is then disconnected.

The master oscillator output jack on the 40A meter is connected to the meter input jack of the WWV receiver with a coaxial cord. The master oscillator is adjusted so that the same reading as before is obtained on the WWV receiver S meter. The WWV antenna is reconnected. An audible heterodyne will be heard in the WWV receiver speaker. The master oscillator adjustment is tuned in the direction which causes the heterodyne to decrease in frequency, until no more than one heterodyne burst occurs each 3 to 5 seconds. This can also be noted visually by needle excursions on the S meter. The 40A meter is now calibrated and ready for use in setting the stable oscillators.

The RF coupler is now connected to the RF input jack on the 40A meter with a coaxial cord. The transmitter is placed on the air and the coupler and RF adjustment are set to give a reading within limits marked in red on the 40A S meter. The COARSE tuning control on the stable oscillator platter is turned counterclockwise until the capacitor plates are unmeshed. The FINE tuning control is set at midpoint. The COARSE control is then turned slowly clockwise until the deviation meter reads about zero, and a final zeroing adjustment is made with the FINE control. The stable oscillator is now set to give a transmitter output on the assigned frequency within the 45 cycle limits of the test gear. However, since a difference in transmitted frequencies of less than 50 cycles is undesirable, it is our practice to then readjust the stable oscillator to give a transmitter frequency 50 cycles away from the assigned frequency, using alternate + and - settings for adjacent transmitters.

While adjustment of the stable oscillator is in progress, the S meter on the WWV receiver is observed. Any instability indicates drifting in the 40A meter, which is then recalibrated.

MODULATION PHASE EQUALIZATION

General

To improve the reception of the same signal in an FM receiver, located in an equal signal area from two adjacent FM transmitters, without disabling one transmitter or by the use of directional mobile antennas, it was necessary to determine the factors that affected the signal reception.

Laboratory and Field Tests

With a setup as shown in figure 15, and by subsequent field tests, the results as indicated below were observed for different conditions. These tests involved the reception of a common signal modulating two FM transmitters by an FM receiver located in an area of equal signal strength where the "capture" of the receiver by one transmitter would not occur. In these tests, the transit or delay time of the modulating signals to the transmitters could be varied and, in addition, the "poling" or the "normal or reversed" position of the wires from one of the lines to one of the transmitters could be changed. Tests were made for various conditions of "delay" and "poling" for both single frequency tone and speech modulating signals to the transmitters.

Tests Using Speech Modulation

Using male voices for speech modulation with the same setup as shown in figure 15, the following results were obtained.

a) When the two lines had the same transit time and both lines were poled alike, the received signal was at a maximum and free of distortion.

b) When the two lines had the same transit time and one of the lines was reversed, almost complete cancellation of the received signal was obtained.

c) When the difference in delay between the two lines was about 250 microseconds or more, the effect of poling one of the lines, normal or reversed, did not produce any appreciable difference in the signal reception for either position. A large amount of distortion was, however, observed in the received signal.

This last condition is normally encountered when telephone lines are used to connect transmitters on a ribbon type mobile radio system and the transit time for each line is not made alike for the voice frequency band width.

It should be emphasized at this point that the delay or phase differences discussed above refers only to the modulating signal which was applied to the two transmitters and the results are independent of how closely the carrier frequencies of the two FM transmitters are held together. As pointed out in the first part of this paper, differences between the carrier frequencies in the two transmitters may result in audible heterodyning, and if this frequency lies in the voice band, it is much more annoying and reduces the intelligibility of the wanted signal more than if it is held by stable oscillators to within 50-150 cycles.

Tests with Single Frequency Tone Modulation

Figure 16 indicates the results of field

tests using a 1,000 cycle tone as the modulating signal to both transmitters. The test conditions and results observed were as follows:

a) When the signals to the two transmitters had the same transit time and the two lines were both poled alike, the resulting signal in the receiver was free from distortion and at its maximum volume. This ignores the effect of the FM carriers cancelling when they are in phase opposition and causing the squelch circuit to operate.

b) When the line to one transmitter was reversed, the resultant signal in the receiver cancelled out. In this case the delay or transit time was the same to both transmitters.

These tests indicated the effect of polarity of the lines to the two transmitters when the delay or transit time for both lines was the same.

Tests under the same conditions as above, but with the lines adjusted to have a difference in transit time of 1,000 microseconds between them, produced the same results as observed in (a) above when the lines had equal delay. This was due to the fact that 1,000 microseconds time delay caused a phase shift of 360° for 1,000 cycles and the tones to the transmitters were again in phase.

With the lines adjusted to a difference of 500 microseconds between them, a 180° phase shift at 1,000 cycles resulted, and with both lines poled alike, the received signal was at a minimum and, conversely, with one of the lines reversed, the signal was at a maximum.

With the lines adjusted to a difference of 250 microseconds between them, which produced a phase shift of 90° for 1,000 cycles, then the poling of the lines made no difference in reception.

Differences in transit time between the two lines caused different amounts of phase shift for various frequencies as indicated in figure 17. These tests indicated that the best reception was observed when the modulating signal to the FM transmitters were in phase and the lines are both poled alike.

It can be seen with single frequency tone modulation that for certain discreet increments of delay, the tone is again in phase and exactly the same as it was initially. For voice modulation, this is not so as any differences in delay to the two transmitters results in different information to them at any particular instant of time.

Application of Delay Equalization to a Working System

As a result of the above tests, it was decided to apply these principles to a working system. It should be noted that it was not necessary to have the delay over the voice band linear with respect to frequency, but only to have equal delay at all frequencies in the voice band if the resultant signal in the interaction areas was to be entirely free from distortion. Field tests indicated that some compromise was possible with these severe requirements. This was important, as the various types of telephone lines in use had a wide variety of delay characteristics over the voice frequency band, and it would be impractical to build out all of these lines for equal transit times at all frequencies. Physical telephone lines, the R carrier on microwave, and the K, N and ON cable carrier telephone lines were being used to connect the FM transmitters.

Method of Delay Equalization

It was decided for practical purposes to hold the transit time differences between the two telephone lines to at least a value that would result in less than a 45° phase shift for the voice frequency band from 500 to 1500 cycles. From figure 17 it can be seen that this permits 250 microseconds differences at 500 cycles but only about 83 microseconds at 1500 cycles. With no more than these differences between the two modulation signals, there would not be cancellation at any frequency but there would be noticeable distortion observed in the received signal with both lines poled alike. However, with a reversal of one line there would be a large amount of cancellation within the voice band. Figure 18 indicates the results obtained in field application of equalizing the delay of the telephone lines.

Delay of the Various Telephone Lines

It was necessary to determine whether the various telephone lines could be brought within the above differences by merely adding the same amount of delay to all of the frequencies in the voice band, or whether a varying amount of delay would be required for the various frequencies. The absolute delay at various frequencies on a telephone line cannot be measured unless both ends of the circuit are available at the same point for connection to a delay measuring set. As this was not the case, other methods had to be used.

Method Used to Determine if Two Telephone Lines Could be Delay Equalized

The relative delay for one frequency with respect to other frequencies for a telephone line can be determined with delay measuring sets at both ends of the line. Figure 19 shows the relative delay measured for two telephone lines used to feed two adjacent transmitters. In this case, a fixed amount of delay for all frequencies within the voice band can be added to one or both of the lines and the resultant transit time for both lines would have reasonable differences in delay between 500 and 1500 cycles. This was determined by superimposing one of the envelope delay curves

on the other and moving it up or down (keeping the edges of the paper parallel). This is equivalent to adding fixed amounts of flat delay to one of the lines until the minimum differences in delay is obtained between the curves from 500 to 1500 cycles. As the remaining differences in microseconds did not exceed an equivalent 45° phase shift for any of the frequencies being considered, the lines could be successfully equalized.

Determination of the Maximum Delay to Build Out All of Lines

Figure 20 shows the operation of the radio system and it will be seen that one of the transmitters is located at the bridging point, which was assumed to have zero delay in the voice path to the transmitter.

The other telephone lines had different transit times which were dependent on the length and type of facilities that were used. The transit time at 1,000 cycles was calculated by using the tables in figure 21. The largest value of transit time at 1,000 cycles was increased by 500 microseconds, to take care of possible differences between the actual line delay and the calculated values. As the largest calculated value was about 3,500 microseconds, 4,000 microseconds was selected to which all lines would be built out. Having determined the calculated values for each telephone line, it was possible to order various delay networks with values that could be used to delay equalize all of the telephone lines. The schematic diagram and photographs of these networks are shown in figures 22, 23, and 24. The electrical characteristics of the networks were specified as follows:

Attenuation between 600 ohms will be not more than:
0.5 db/millisecond at 1,000 cycles
1.1 db/millisecond at 3,000 cycles

Envelope Delay ± 1% for the range 500-3,000 cycles. These networks were obtained from the EPSCO Company, Boston Massachusetts

Use of Fixed Delay at the Bridge Transmitter

A fixed 4,000 microsecond delay network was installed in the line to transmitter A from the bridge as shown in figure 20. However, as seen in figure 19, the delay shape of the telephone lines, because of the cutoff of the lower and higher frequencies, was tipped up at both ends of the voice band. As the 4,000 microsecond network was practically flat for the 300 to 3,000 cycle band, it was necessary to add a repeating coil and a special delay network to tip up both ends of the frequency band. The delay shape of this network was then similar to the telephone lines. The absolute delay was measured at 1,000 cycles and was 4,000 microseconds.

If a method could be devised to adjust the adjacent lines to the same delay as the above network within a 45° phase difference tolerance in the 500 to 1,500 cycle band, it would be unnecessary to determine the absolute delay value of the adjacent lines. Figure 18 indicates that when the two lines are equal in transit time and one line is reversed to provide approximately a 180° phase difference, almost complete cancellation of the signal from the two transmitters takes place in the receiver. This principle was used to adjust the delay in each line to its adjacent line, starting with the line containing the fixed 4,000 microsecond delay network.

Adjustment of the Line Adjacent to Bridge Transmitter

The delay adjustment of the adjacent line with respect to the fixed delay in the bridge transmitter was made in the field as follows: A mobile receiver was located in the area where the signals from the two transmitters were equal. A delay of about 200 microseconds less than the calculated value required to equal the 4,000 microsecond network was added to telephone line B. In addition a delay network, adjustable in steps of 20 microseconds over the amount calculated, was arranged to be switched in as required. Also, a switch to reverse the line to the transmitter was placed in the circuit. The two transmitters were then modulated by the same voice signal. A voice recording was used, as this provided uniform modulation even after many repetitions (which is not true for people) and generated all frequencies in the voice band to permit the best compromise setting of the delay in line B for voice transmission. Delay was added in 20 microsecond steps in line B and the signal was monitored at the mobile receiver with the both lines normal and with one line reversed. As the delay was increased, the amount actually required to best equalize line B to the value of line A was reached. This was determined when maximum cancellation was obtained with one line reversed. This method was found to be practical and it was used to delay equalize all lines with respect to each other starting with those adjacent to the line having the fixed network and proceeding until all lines were thus adjusted. Conversely, when the line was normal, the signal was at its maximum volume. The point of maximum signal strength was difficult to observe in the field as the increase was not too obvious, nor was it always obtained. On the other hand, the cancellation or reduction of signal, generally in the order of 6 db or more, was easily observed as the delay of the two lines became equal and provided positive identification of this point.

Types of Facilities Usable

After the telephone lines were delay equalized, it was necessary to keep the poling alike as a reversal of one line would result in

poor reception in the interaction area. It was found that telephone lines using microwave or K or L cable carrier systems were not suitable for use in this service as the voice shifted in phase after a period of time even when they were in phase at the transmitting end of the lines. This was due to the fact that the carrier used for modulation is not the same as the carrier used for demodulation, therefore, differences in frequency between the two caused phase shifting of the voice. On the other hand, the ON and N systems which used transmitted carrier at the demodulating end resulted in the phase angle of the voice remaining fixed with the transmitting end. These two carrier systems and physical voice circuits were found suitable for use as land lines in this mobile radio system.

On this basis it appears that even if microwave circuits were used to connect together all the transmitters and they all had the same transit time, it would also be necessary to transmit the modulating carrier to be used for demodulating or use some means of tieing the two carriers together at the same frequencies as any change or drift would eventually result in phase shifting of the voice which would be equivalent to reversing one of the lines.

Temperature Effects

It is known that temperature changes affect the absolute delay of the telephone facilities. However, in two years of service temperature effects have all proved to be in the same direction for the different seasons and consequently have not made enough difference to noticeably increase the distortion in the interaction areas. Thus, it has been unnecessary to re-equalize the line facilities due to seasonal temperature changes.

Conclusion

Intelligible reception in interaction areas of a ribbon type mobile system can be provided by the above method without disabling a transmitter or operating a directional type mobile antenna. Field results indicate that this method produces a signal that is understandable even with distortion in the interaction areas, whereas previously, these sections had large amounts of distortion and were unintelligible in many places. In practice, the two telephone lines should be equalized for delay at least from about 500 to 1,500 cycles. The difference in delay may then be as high as 80 microseconds at the higher frequencies and 250 microseconds at the lower frequencies. However, the quality of the received signal is considerably improved where the delay of each frequency in the two lines is, or is nearly the same and for a band width beyond 1,500 cycles. Flat delay equalization was found to be practical in most cases, where judicious selection of telephone lines is made. In some cases, the addition of a simple transformer sufficed. Differences in delay caused by temperature variations in our plant appears to be sufficiently close to wash out.

The use of stable oscillators on the FM transmitters was found to be very desirable in further improving intelligibility of reception of speech transmission on receivers in the "interaction" areas.

References

Bell Laboratories Record, November 1956 - Reducing Distortion in Mobile Radio Systems - W. C. Babcock and R. V. Crawford.

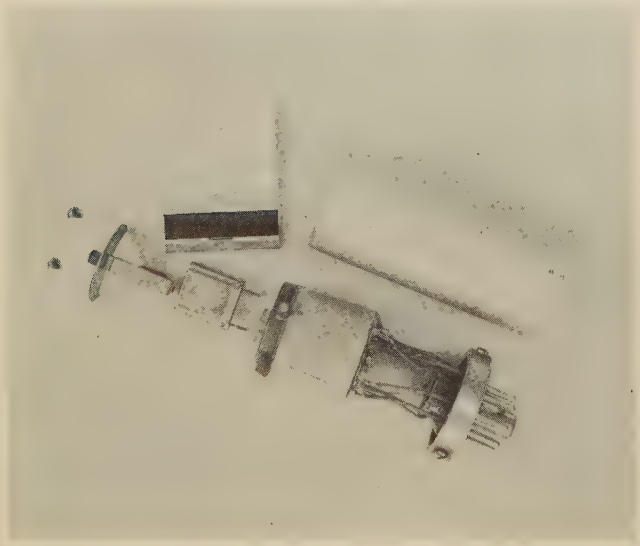
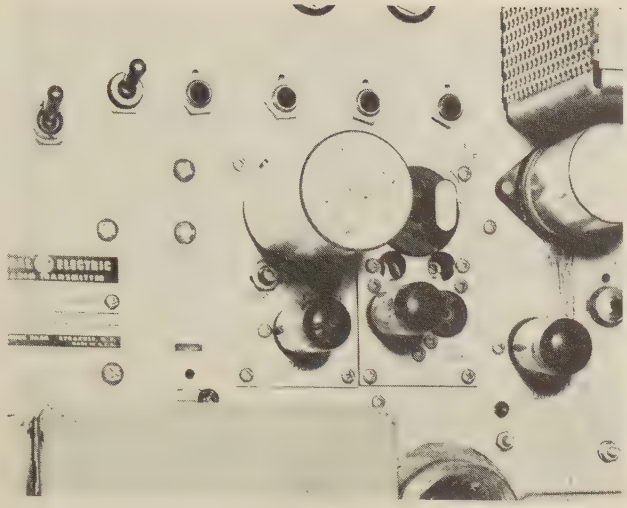
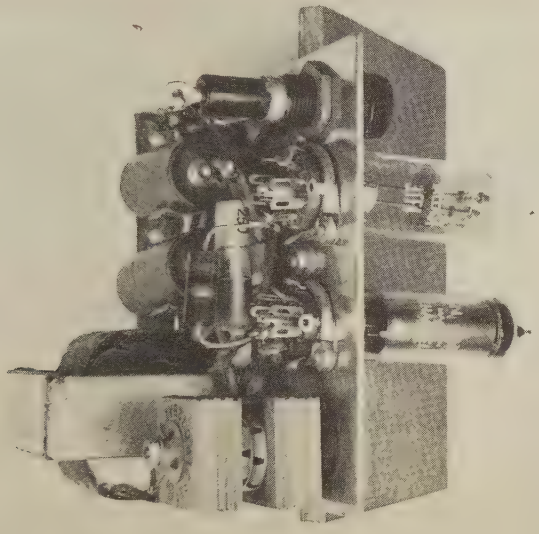
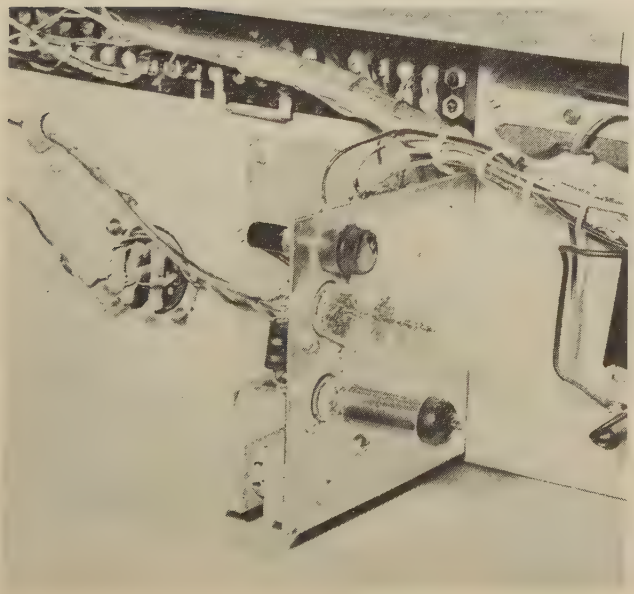


Fig. 1 - Crystal and Oven

Fig. 2 - Stable Oscillator Platter







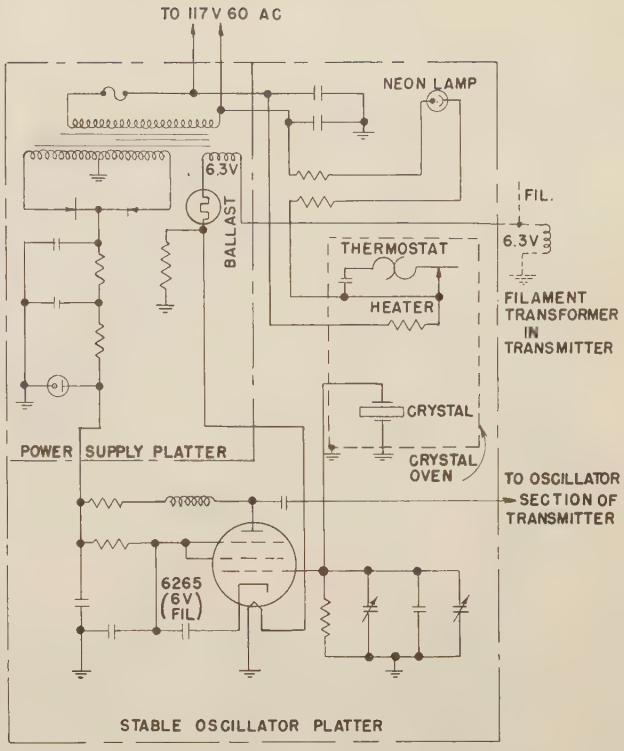


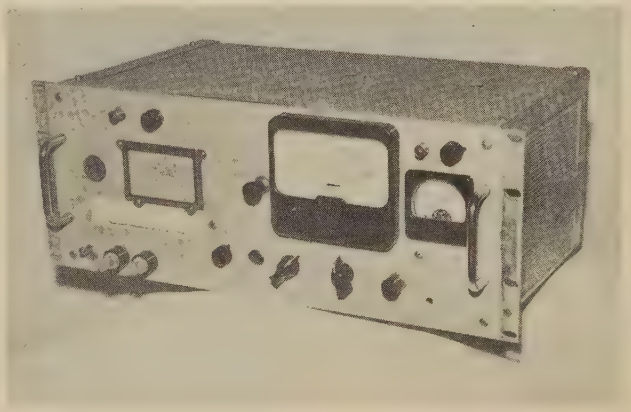
Fig. 6 - Stable Oscillator and Power Supply Circuit (Above)

Fig. 3 - Stable Oscillator Platter Mounted (Upper Left)

Fig. 4 - Power Supply Platter (Center Left)

Fig. 5 - Power Supply Platter Mounted (Lower Left)

Fig. 7 - 40A Frequency Meter (Below)



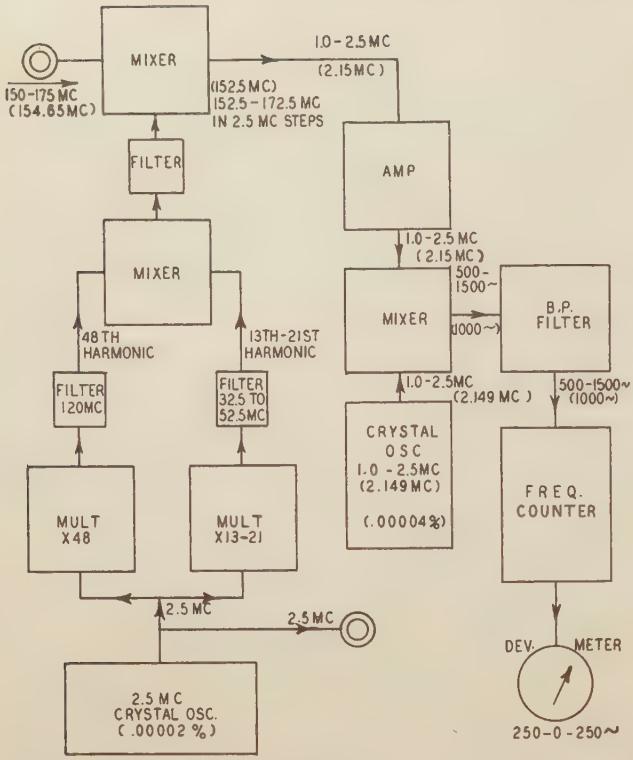


Fig. 8 - 40A Frequency Meter Block Diagram

Fig. 9 - WWV Receiver

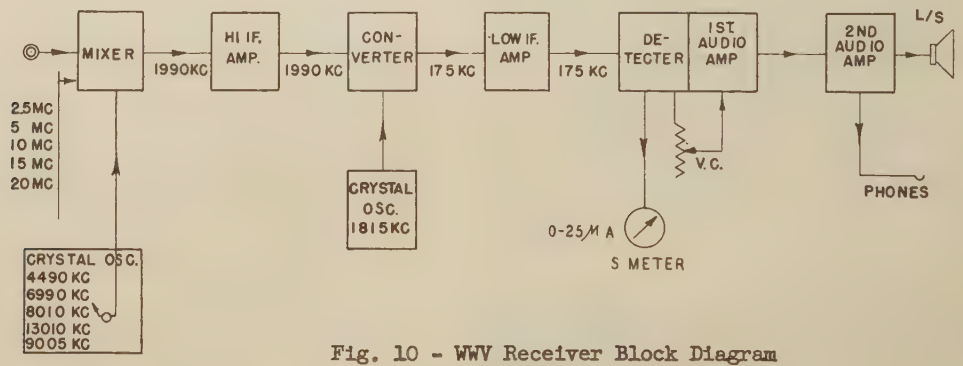


Fig. 10 - WWV Receiver Block Diagram

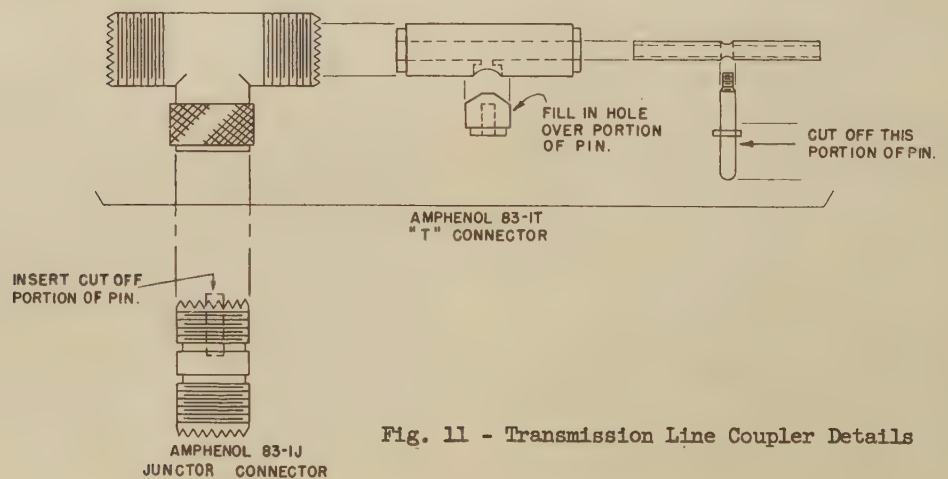


Fig. 11 - Transmission Line Coupler Details

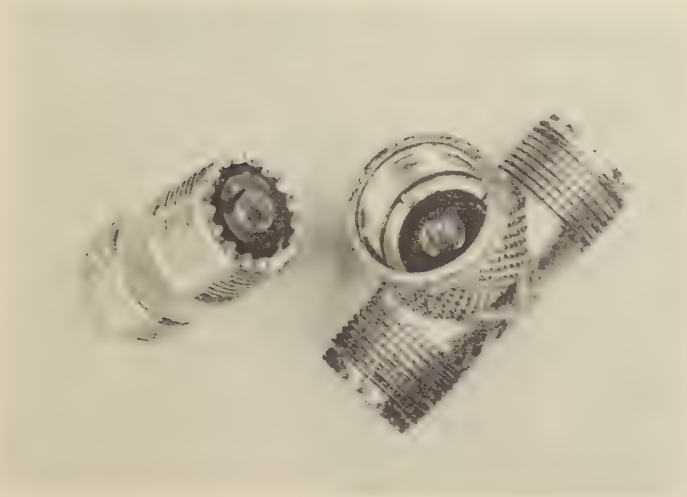


Fig. 12 - Transmission Line Coupler

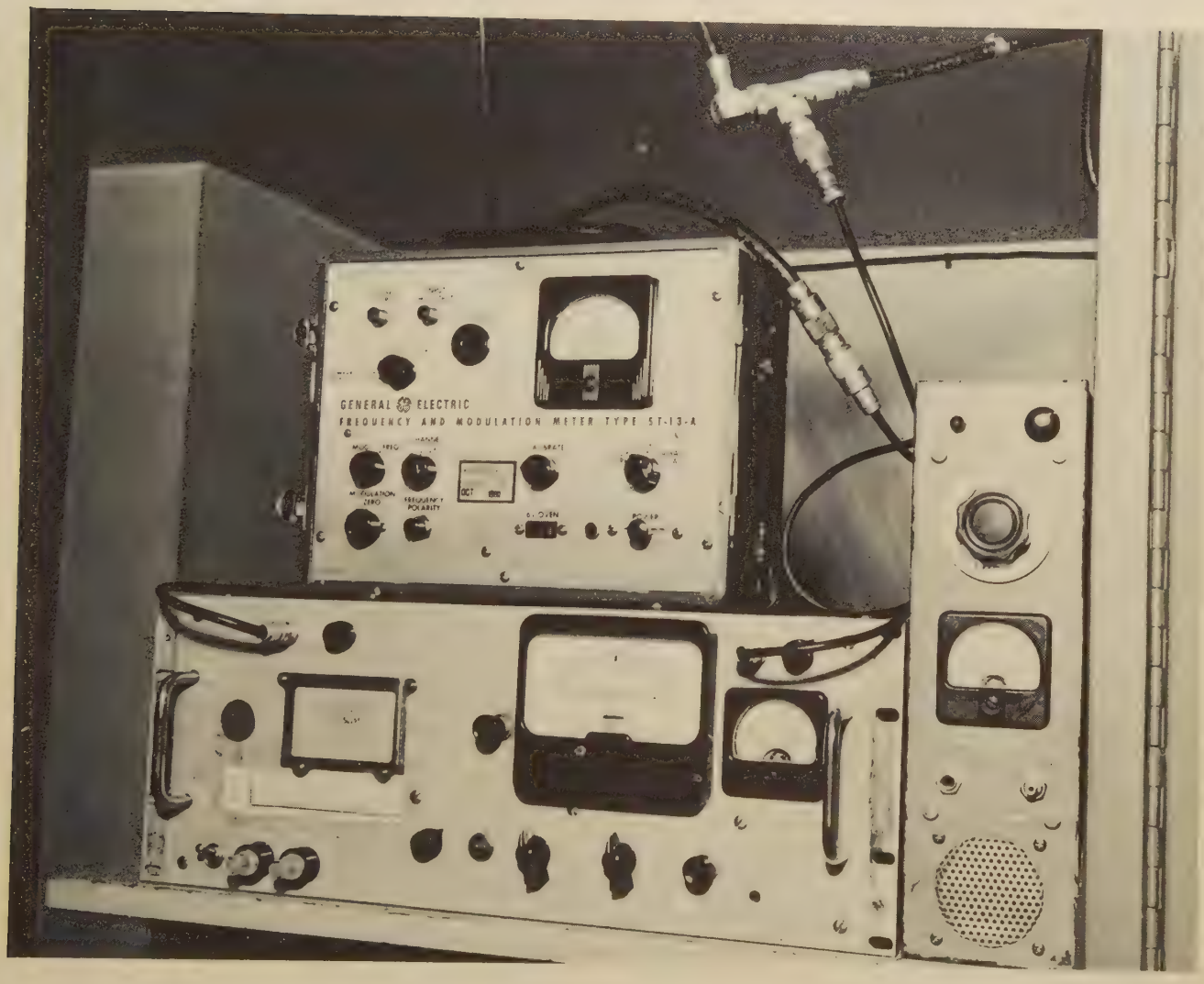


Fig. 13 - Test Equipment Assembly

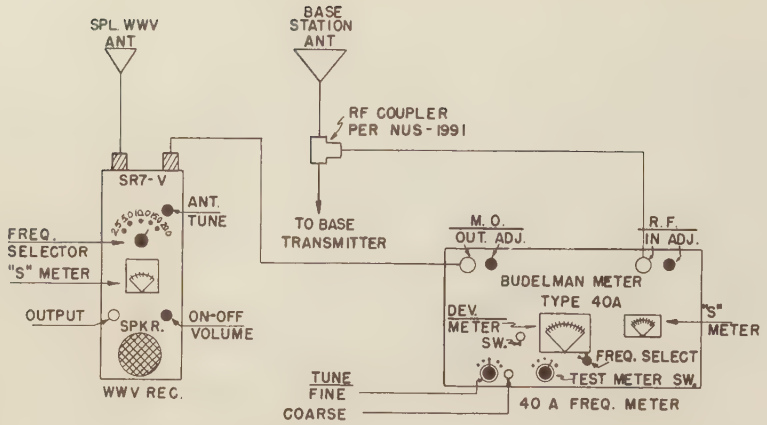


Fig. 14 - Test Equipment Diagram

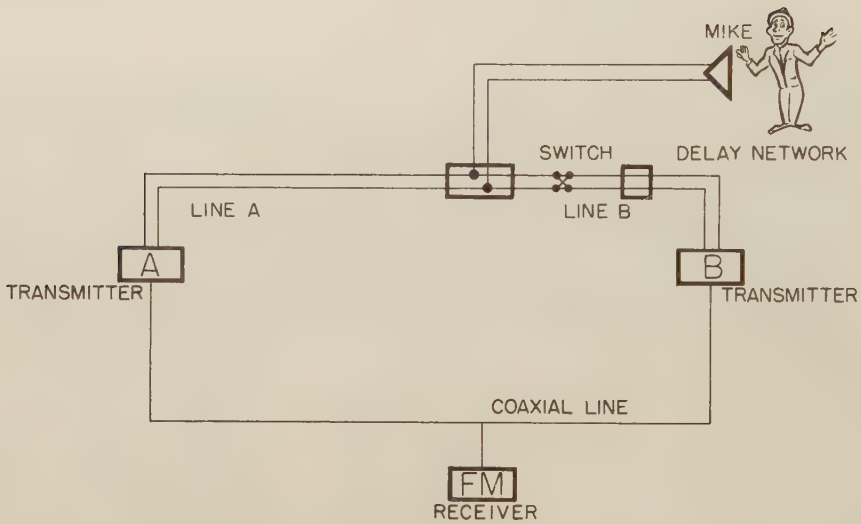


Fig. 15 - Laboratory Demonstration

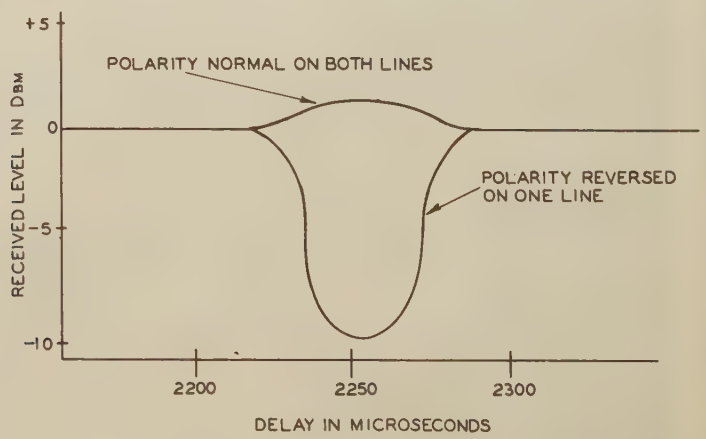
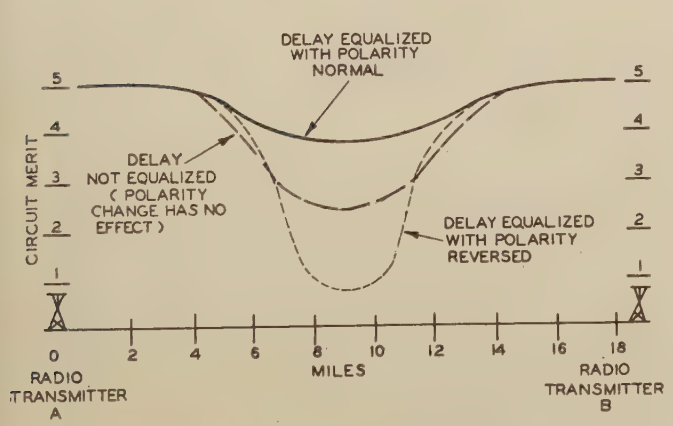


Fig. 16 - Polarity Effect Using 1,000 Cycle Tone

MICROSECOND EQUIVALENT OF PHASE ANGLE FOR VARIOUS FREQUENCIES

Frequency in C. P..S.	360°	180°	90°	45°
100	10000.0	5000.0	2500.0	1250.0
200	5000.0	2500.0	1250.0	625.0
300	3333.0	1666.0	833.0	416.6
400	2500.0	1250.0	625.0	312.5
500	2000.0	1000.0	500.0	250.0
600	1666.0	833.3	416.6	208.0
700	1428.5	714.2	357.1	178.5
800	1250.0	625.0	312.5	156.2
900	1111.1	555.5	277.7	138.8
1000	1000.0	500.0	250.0	125.0
1100	909.0	454.5	227.7	113.6
1200	833.3	416.6	208.3	104.1
1300	769.2	384.6	192.3	96.1
1400	714.2	357.1	178.5	89.2
1500	666.6	333.3	166.6	83.3
1600	625.0	312.5	156.2	78.1
1700	588.2	294.1	147.0	73.5
1800	555.5	277.7	138.8	69.4
1900	526.3	263.1	131.5	65.7
2000	500.0	250.0	125.0	62.5
2100	476.1	238.0	119.0	59.5
2200	454.5	227.2	113.6	56.8
2300	434.7	217.3	108.6	54.3
2400	416.6	208.3	104.1	52.0
2500	400.0	200.0	100.0	50.0
2600	384.6	192.3	96.1	48.0
2700	370.3	185.1	92.5	46.2
2800	357.1	178.5	89.2	44.6
2900	344.8	172.4	86.2	43.1
3000	333.3	166.6	83.3	41.6
3100	322.5	161.2	80.6	40.3
3200	312.5	156.2	78.1	39.0
3300	303.0	151.5	75.7	37.8
3400	294.1	147.0	73.5	36.7
3500	285.7	142.8	71.4	35.7
3600	277.7	138.8	69.4	34.2
3700	270.2	135.1	67.5	33.7
3800	263.1	131.5	65.7	32.8
3900	256.4	128.2	64.1	32.0
4000	250.0	125.0	62.5	31.2

Fig. 17 - Microsecond Equivalent of Phase Angle for Various Frequencies



Circuit Merit	
CM5	Perfectly readable, negligible noise.
CM4	Perfectly readable but with noticeable noise.
CM3	Readable with few syllables missing, requires occasional repetition (commercial).
CM2	Readable with difficulty, requires frequent repetition (noncommercial).
CM1	Unusable, presence of speech barely discernible.

Fig. 18 - Effect of Polarity on Voice Transmission

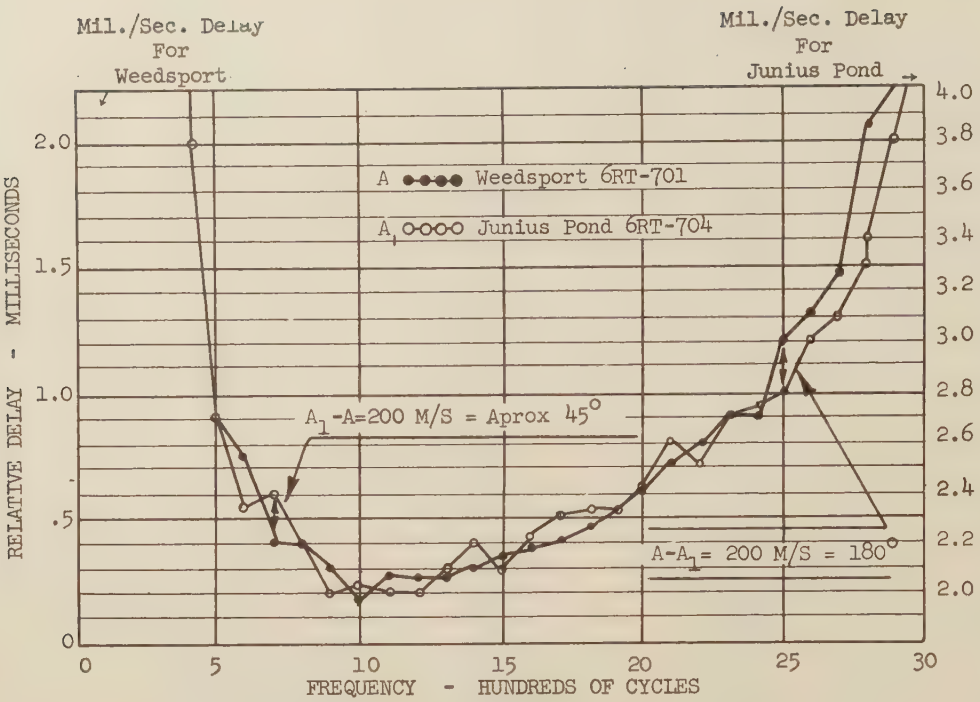

Mil./Sec. Delay For Weedsport
Mil./Sec. Delay For Junius Pond
RELATIVE DELAY - MILLISECONDS
A ●●●● Weedsport 6RT-701
A1 o-o-o-o Junius Pond 6RT-704
A1-A=200 M/S = Aprox 45°
A-A1= 200 M/S = 180°
FREQUENCY - HUNDREDS OF CYCLES

Fig. 19 - Comparison of the Delay of Two Adjacent Telephone Lines

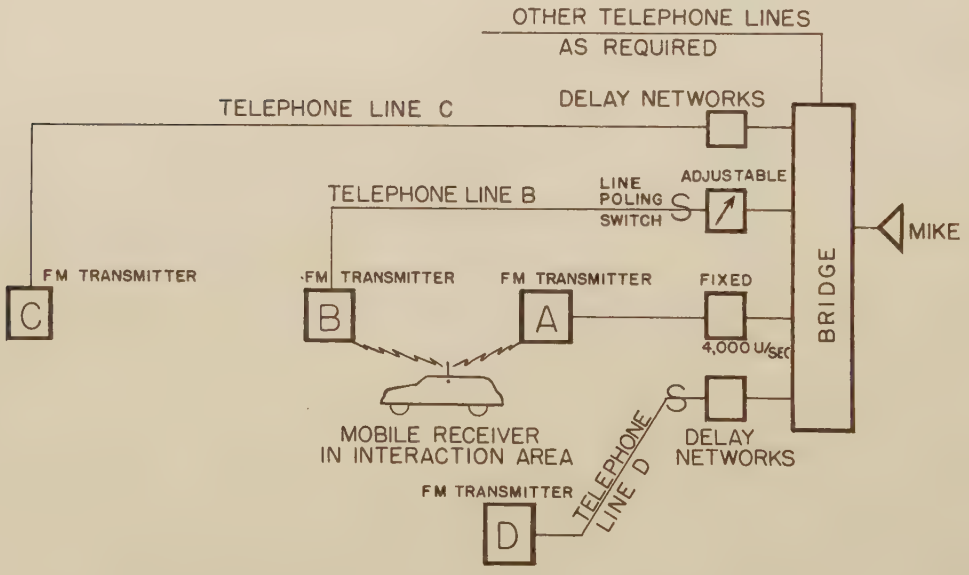


Fig. 20 - Operation of Ribbon Mobile Type Radio System

TABLE OF DELAY VALUES FOR COMMON TELEPHONE FACILITIES

Toll Facilities: (At 1000 CPS, 68°F in Microseconds/Loop Mile)

Gauge	Side					Phantom			
	Loading	Prop. Const. Per Loop Mile		Phase Delay		Loading	Prop. Const. Per Loop Mile		Phase Delay
	NL	.1249+j	.1338	21.31		NL	.1106+j	.1219	19.41
	B-88	.0322+j	.6186	98.50		B-50	.0273+j	.5933	94.47
	H-31	.0643+j	.2693	42.88		H-18	.0529+j	.2642	42.07
19	H-44	.0561+j	.3138	49.97	19	H-25	.0466+j	.3047	48.52
	H-88	.0418+j	.4388	69.87		H-50	.0351+j	.4228	67.32
	H-172	.0323+j	.6085	96.89		H-63	.0331+j	.4712	75.03
	H-174	.0331+j	.6107	97.25		H-106	.0269+j	.6129	97.60
	NL	.0842+j	.0974	15.08		NL	.0746+j	.0890	14.20
	B-88	.0185+j	.6181	98.42		B-50	.0157+j	.5929	94.41
	H-31	.0334+j	.2638	42.01		H-18	.0273+j	.2604	41.46
16	H-44	.0296+j	.3134	49.90	16	H-25	.0243+j	.3022	48.12
	H-88	.0224+j	.4374	69.65		H-50	.0189+j	.4218	67.16
	H-172	.0183+j	.6082	96.85		H-63	.0185+j	.4705	74.92
	H-174	.0191+j	.6102	97.17		H-106	.0156+j	.6126	97.55
13	NL	.0538+j	.0740	11.78	13	NL	.0442+j	.0710	11.30
10	NL	.0342+j	.0600	9.55	-				

CARRIER AND REPEATERS

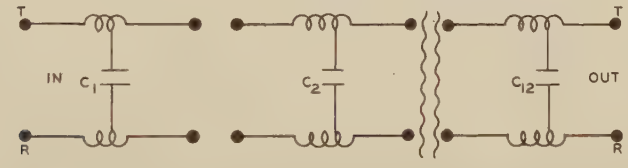
Carrier Facilities:

	(1000 Cycles)	(1200 Cycles)
N-1	9.3 Microsec/Mile/Approx.	1120 Microsec/Pr. of Terminals
ON	" " "	
O	" " "	
K-1	8.32 " " "	
K-2	7.72 " " "	940

Repeaters:

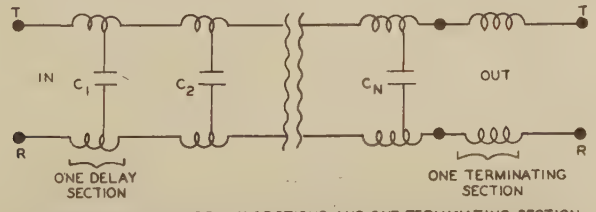
44A1	75	"
22A1	125	"
V1	14	"
V3	41	"

Fig. 21 - Table of Delay Values for Common Telephone Facilities



1 MILLISECOND ADJUSTABLE UNIT HAS 12 DELAY SECTIONS AND NO TERMINATING SECTION

1 MILLISECOND UNIT WITH SWITCHES HAS 12 SWITCHES CONNECTED LIKE 200 MICROSECOND UNIT



8 MILLISECONDS HAS 96 DELAY SECTIONS AND ONE TERMINATING SECTION

4 MILLISECONDS HAS 48 DELAY SECTIONS AND ONE TERMINATING SECTION

1 MILLISECONDS HAS 12 DELAY SECTIONS AND ONE TERMINATING SECTION

Fig. 22 - Schematic of Delay Networks

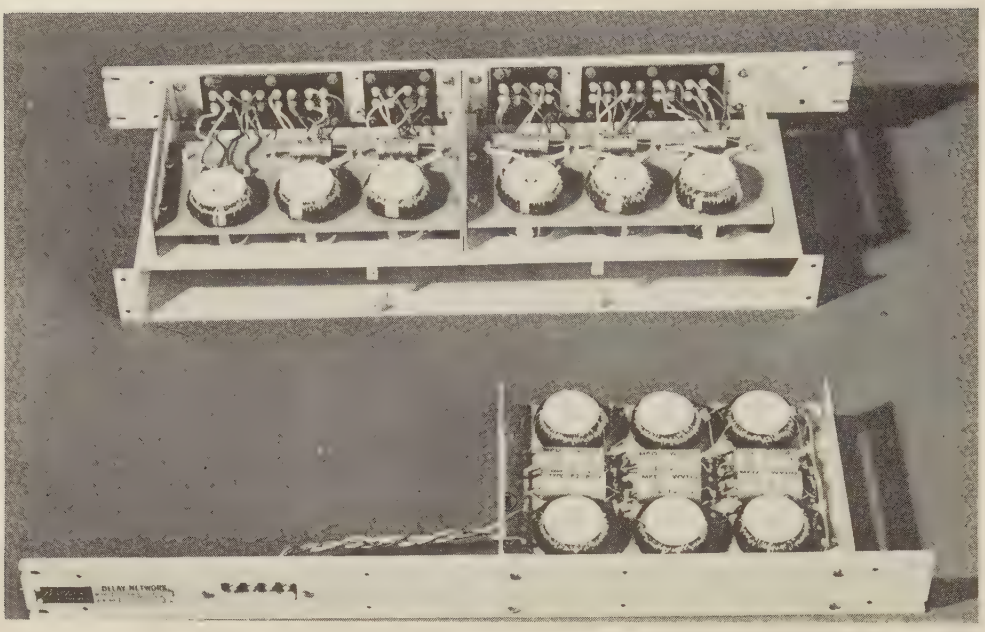


Fig. 23 - Photograph of Delay Networks

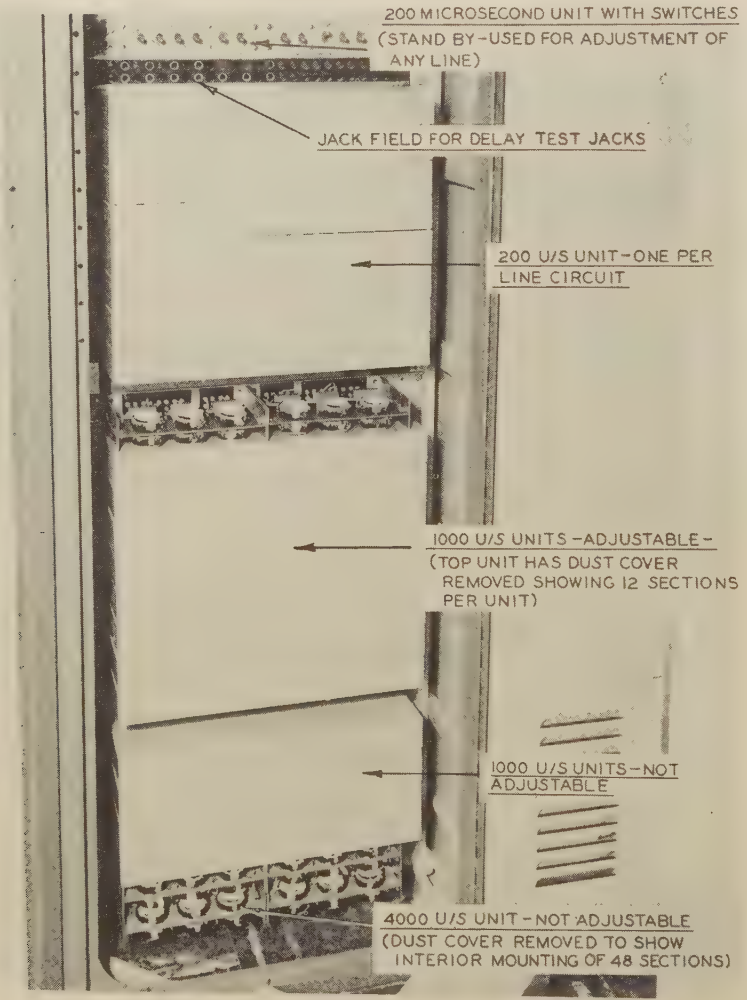

200 MICROSECOND UNIT WITH SWITCHES
(STAND BY-USED FOR ADJUSTMENT OF ANY LINE)
JACK FIELD FOR DELAY TEST JACKS
200 U/S UNIT-ONE PER LINE CIRCUIT
1000 U/S UNITS-ADJUSTABLE-
(TOP UNIT HAS DUST COVER REMOVED SHOWING 12 SECTIONS PER UNIT)
1000 U/S UNITS-NOT ADJUSTABLE
4000 U/S UNIT-NOT ADJUSTABLE
(DUST COVER REMOVED TO SHOW INTERIOR MOUNTING OF 48 SECTIONS)

Fig. 24 - Photograph of Delay Networks

COMPARISON OF WIDEBAND AND NARROWBAND MOBILE CIRCUITS AT 150 MEGACYCLES

H. W. Nylund
Bell Telephone Laboratories, Incorporated
Murray Hill, New Jersey

Summary

This paper gives results of laboratory tests made to evaluate the relative transmission performances of wideband and narrowband mobile radio telephone circuits in the 150 mc band. It was found that the effects of changing from wideband to narrowband operation are dependent upon RF impulse noise, frequency errors and RF signal levels. Under ideal conditions the change to narrowband may improve circuit performance; but where impulse noise interference is severe circuit performance will be seriously degraded unless rather large compensating changes can be made in such factors as RF signal power or average modulating levels.

Introduction

Split channel operation with 30 kc channel spacing is now a familiar fact in the 150 mc mobile band. F. C. C. rules require that by 1963 all transmitters must meet new narrowband standards as to frequency stability, maximum deviation and control of extra-band modulation. While the rules do not specify receiver characteristics, split channel operation has also made it necessary for manufacturers to provide narrowband receivers with improved stability and selectivity. The principal features of the old wideband and the new narrowband equipments are listed in Fig. 1. As shown, a wideband mobile circuit in our Bell System operation has ±.002% frequency stability; the transmitter uses ±15 kc maximum deviation and the receiver has a nose bandwidth of about 30 kc at the 6 db points. In the narrowband circuit frequency stability is tightened to ±.0005%, maximum deviation is reduced to ±5 kc, and receiver nose bandwidth is narrowed to about 12 kc.

In laying out new systems or converting old ones it is necessary to know how the transmission performance of the circuit will be affected by the change from wideband to narrowband operation. The possible transmission effects can be divided into two classes. Multichannel effects may, of course, result from the interaction of two or more channels to cause interference, intermodulation, etc. Other effects upon signal-to-noise ratios, speech levels and distortion may occur in a single isolated channel. We have recently made a number of laboratory tests to evaluate this second class--single-channel effects--by comparing the transmission performances of a wideband and a narrowband circuit at a frequency of about 153 mc.

Test Setup and Methods

The setup used for these tests is shown in Fig. 2. Recorded speech was supplied from the tape reproducer, sent through the control terminal and applied to the audio input of the transmitter. The transmitter was arranged so that it could readily be switched from the wideband condition with ±15 kc maximum swing to the narrowband condition of ±5 kc swing. A small portion of the transmitter RF output was applied either to attenuator A and wideband receiver A, or to attenuator B and narrowband receiver B. The audio outputs of these two receivers were similarly arranged so that either could be switched to the vehicular control unit and handset used by the listening observer; and a test set at this point permitted measurement of speech and noise levels.

Going back to the transmitter, its audio input circuits were adjusted so that +7 dbm of 1000 cps test tone applied at the control terminal caused ±10 kc swing in the wideband condition, or ±3.3 kc swing for the narrowband.* The vogad in the control terminal held applied speech levels to a very small range about 0 VU. This put the wideband and narrowband circuits on an equal footing as far as modulation was concerned, since in both cases there was a difference of 7 between the dbm level of test tone for full system modulation, and the VU level of the applied speech.

One other feature of this test arrangement should be mentioned. In some tests a noise generator as shown in Fig. 2 was used to supply RF impulse noise, which was adjusted in level by attenuator C and applied to the receiver input along with the wanted signal.

In a typical comparison test with this setup the audio gains of the two receivers were first adjusted so that, with a strong RF signal, both circuits gave the same speech levels at the observer's handset. Attenuator A was next adjusted to give the wideband receiver an RF signal which would produce a measured speech-to-noise ratio of 12 db at the handset, and this setting of attenuator A was left unchanged throughout the test. Then rapid switches were made so that the observer listened alternately to the wideband and

---

*With this lineup condition, the same level of 1500 cps tone (which most nearly represents the speech wave in its effect upon a phase modulator) would cause maximum deviation of ±15 kc in the wideband transmitter, or ±5 kc in the narrowband.

narrowband circuits, and he adjusted attenuator B until he judged that the two circuits were equal in transmission quality. The difference in settings between attenuators A and B was then taken as a measure of the comparative performance of the wideband and narrowband circuits, in terms of the allowable path loss for equivalent transmission quality.

Relative Performance of Wide and Narrowband Circuits

Tests of this kind were made with several different receivers, with and without impulse noise injected into the RF circuit. The character of the speech was varied, using both male and female talkers on the tape. While the test procedure just outlined involved an audio speech-to-noise ratio of 12 db, which gives a barely-commercial circuit, other tests were made with a better circuit having a 22 db speech-to-noise ratio.

Some of the results of these tests are shown in Fig. 3. In this plot the reference condition, shown by the horizontal line at zero on the vertical scale, is the path loss over which the wideband circuit would give a 12 db speech-to-noise ratio. The vertical scale indicates the amount by which this path loss had to be increased or decreased for the narrowband circuit to give equivalent transmission quality. The horizontal scale shows various RF noise conditions. Where there was no impulse noise and the receiver performance was controlled by set noise, some observers judged that the narrowband circuit would tolerate 5 db more path loss than the wideband. Some judged the two as equal, and the average advantage found for the narrowband circuit was 2.3 db.

When impulse noise interference was introduced the narrowband circuit became poorer than the wideband. With a noise peak voltage at the receiver of 60 db above 1 μv per mc, on the average the narrowband circuit required about 13 db less path loss than the wideband for equal performance, with the range for different observers extending from 8 to 17 db. This 13 db average difference might of course be expressed in other terms than path loss. It would be equally accurate to say that the narrowband receiver required 13 db more RF signal than the wideband to give equal transmission performance, or that the narrowband transmitter would need 13 db more power to cover the same geographical range; but in many cases such power increases would not be permissible under F. C. C. rules.

With impulse noise at levels higher than 60 db above 1 μv per mc, the effect remained nearly constant. It should be mentioned that these were on-frequency tests: that is, the RF signal frequency was kept centered on the receiver discriminator within a few hundred cycles.

Noise Characteristics of Receivers

One obvious reason for the above-outlined differences in transmission between wide and narrowband circuits is, of course, the different deviations used at the transmitters. Another contributor was found in the noise performance of the receivers.

For two receivers reasonably typical of those tested, Fig. 4 shows how their measured audio outputs of speech and noise varied with changes in RF signal input. The audio gains of these receivers were adjusted so that the two upper curves, for speech, coincide for all except quite weak signals. The two lowest curves (3) and (4) show the familiar set noise quieting characteristics--the solid line for the narrowband receiver, the dashed line for the wideband. When fairly severe RF impulse noise interference was added, the audio noise output was increased as shown by the two curves (5) and (6) in the middle of the figure.

The vertical distance between the speech and noise curves, as between (1) and (3), is of course equal to the speech-to-noise ratio of the circuit. This has been plotted in Fig. 5. Curves (1) and (2) show speech-to-noise ratios measured on the narrowband and wideband circuits with no impulse noise present. With a marginally-weak RF signal of about -144 dbw the narrowband circuit, curve (1), gave the better S/N ratio by a few db. Curves (3) and (4) show the situation with fairly severe impulse noise interference. On the wideband circuit an RF signal of about -138 dbw gave an S/N ratio of 12 db. As shown by A in the figure, with the same RF signal the S/N ratio on the narrowband circuit was at least 6 db poorer. Or, as shown by B, to restore the 12 db S/N ratio the RF signal input to the receiver on the narrowband circuit had to be increased about 12 db, which agrees quite closely with the average results obtained in the direct comparison tests discussed earlier.

So far we have compared the wideband and narrowband circuits only in the marginal-signal region with an S/N ratio of 12 db. This is a usable circuit, but a rather noisy one. Fig. 5 can also be used to indicate the performance of better circuits. An S/N ratio of 22 db gives a fairly good circuit, although noise is still noticeable. In this region of performance the curves show that the narrowband circuit was some 3 to 7 db poorer than the wideband in S/N ratio, depending on the absence or presence of impulse noise; and with impulse noise present the narrowband circuit required about 7 db more RF signal than the wideband to maintain the same S/N ratio. While Fig. 5 shows results obtained with only one pair of receivers, these comparative performance values are within 1 or 2 db of the averages of all tests made with two wideband and three narrowband receivers.

Transmitter Modulation Levels

These results indicate that when a wideband circuit is changed to narrowband, for fringe-area operation the audio noise performance should not be much affected in situations where RF impulse noise is low. (This of course is exclusive of any multichannel interference effects, and it also assumes on-frequency operation: the effects of frequency errors will be mentioned later.) Where impulse noise is a serious factor, however, the change to narrowband may reduce audio speech-to-noise ratios by some 6 or 7 db. In our Bell System service this is enough to make an originally-commercial circuit quite unusable. Most disturbing is the fact that if we were to recover this 6 or 7 db of S/N ratio solely by raising transmitter power or decreasing the allowable path loss for fringe operation, these factors would have to be changed as much as 12 or 13 db.

Fortunately, we have available another way of improving the situation, at least in base-to-car transmission, by raising average modulation levels. Tests which were made on a narrowband circuit have shown that the modulating margin* at the base transmitter can be reduced to a value as low as 4 without causing excessive distortion. This is 3 db higher average modulation than we have used in any of our wideband systems, and 7 db higher than that used in many cases. By raising modulating levels when we change to narrowband, therefore, we can gain 3 to 7 db in S/N ratio at the expense of only slight speech distortion, and this should considerably reduce or in many cases eliminate the degradations previously discussed.

This method of meeting the situation is of course not applicable where the old wideband circuit already uses maximum modulation levels. Vehicular transmitters in our wideband systems are modulated up to the tolerable limit by strong talkers, and so the distortion resulting from a substantial increase in modulation would cancel any S/N ratio improvement. For the car-to-base direction of transmission, therefore, the only available remedial measures are to put base receivers in quiet locations with low RF noise, or to provide more closely-spaced receivers.

Frequency Error Effects

It is well known that frequency errors--differences between the RF signal and the receiver center frequency--will add to noise problems on mobile circuits. Tests were made with wideband and narrowband receivers to compare the increases in audio noise caused by frequency errors of known magnitude. These tests were made with fairly severe impulse noise,** and at RF signal levels which would provide a fringe-area circuit with 12 db speech-to-noise ratio for on-frequency operation. The results, for two receivers reasonably typical of those tested, are shown in Fig. 6. Comparison of the two curves indicates that for a given increase in audio circuit noise under these conditions, the frequency error in the narrowband circuit must be held to only 1/3 to 1/4 of that allowable in the wideband. Further, we may consider what this effect means in terms of expectable frequency errors. The narrowband stability requirement is ±.0005%, and at 150 mc this corresponds to ±750 cps, or a possible total error of 1.5 kc if the transmitter and receiver are off frequency in opposite directions. From curve (1) this would cause a noise increase of 7 db, which is enough to make an originally-marginal circuit quite unusable for our purposes. With good equipment and good maintenance, of course, one would expect that average frequency errors could be held to less than the 1.5 kc just cited; on the other hand, inaccurate setting or drift in an IF stage or discriminator may add to the total error.

Intermixed Operation

During the period of transition from wideband to narrowband operation there may be situations requiring intermixed operation: that is, use of wideband receivers with narrowband transmitters, or vice versa. In such cases the audio gain of the receiver must of course be readjusted to give a suitable listening level under the modulation conditions actually used in the transmitter; but if this is done the tests indicated that intermixed operation is practicable within certain limitations.

An intermixed circuit which has a narrowband transmitter operating into a wideband receiver may be compared with a regular narrowband circuit. For on-frequency operation the two circuits are identical except for the receiver bandwidth, and the wider bandwidth used in the intermixed circuit may be expected to give it somewhat poorer S/N performance. Test results showed that an intermixed circuit of this kind required some 2 to 5 db less RF path loss (or more RF signal) than the standard narrowband circuit for equivalent performance: the exact amount of the difference depending upon the presence or absence of impulse noise. If this intermixed circuit is compared with a standard wideband circuit, these results indicate that where impulse noise is severe the intermixed arrangement will need on the average about 15 db more RF signal at the receiver to give as good transmission.

Where the intermixed circuit has a wideband transmitter operating into a narrowband receiver it is necessary to consider the possibility of distortion resulting from speech modulation swinging outside of the receiver passband. Tests showed that this did cause objectionable distortion if speech was applied to the wideband transmitter input at too high a level. With a modulating margin of 7, however, distortion was

---

*"Modulating margin" as here used is equal to the dbm level of test tone for full system deviation, minus the VU level of speech applied to the same input point.

**At peak voltage level of 60 db above 1 microvolt, per megacycle of bandwidth.

not serious. Under this condition, tests were made to compare the intermixed circuit performance with that of a standard wideband circuit, for on-frequency operation. The slight distortion present in the intermixed circuit caused considerable variation in the judgments rendered by various observers. At one extreme of the range some observers decided that the intermixed circuit was 6 db poorer than the standard wideband, in terms of required RF signal: at the other extreme, other observers called it 6 db better. The net conclusion was that on the average these two circuits should be called about equal, when all levels of noise and performance were considered.

Impulse Noise Considerations

The test results earlier presented in Figures 3, 4 and 5 showed that ambient RF impulse noise has a critical effect upon the comparative performance of wide and narrowband circuits. In those tests impulse noise was injected into the radio circuits from a locally-constructed noise generator. The peak voltage of this impulse noise was measured with an Empire Devices meter which gave readings in terms of db above one microvolt, per megacycle of bandwidth. In all, tests of this kind were made with several different noise generators, and also with "live" ignition noise derived from cars parked outside the laboratory. These tests, plus studies of field experience, indicated that the interfering effect of 60 dbmc* noise from the generator was reasonably representative of the fairly severe noise interference often found in urban operation, while 80 dbmc noise probably approached the worst conditions we ordinarily encounter in the field.

As an incidental result, this work reminded us of a fact which is not new, but is sometimes overlooked. The peak voltage of an impulse-type noise does not completely describe its disturbing effect upon the audio circuit. To illustrate: Fig. 7 shows the results of speech-to-noise ratio measurements on a narrowband circuit, in the presence of RF ignition noise picked up from two cars. For curve (1) the car motors were idling, and the loss in the noise path was adjusted to give 60 dbmc noise voltage at the receiver input. For curve (2) the motors were speeded up somewhat, and the noise path loss increased to keep 60 dbmc noise at the receiver. It will be noted that there is a spread of up to 8 db in S/N ratio between the two curves; and in terms of the RF signal level needed for a given S/N ratio the spread is even wider. This suggests a need for some way of measuring impulse noise, in the mobile bands, which will be more accurate than peak voltage readings in indicating potential interference effects. The quasi-peak method of measurement would seem to offer possibilities which should be investigated.**

Conclusions

While the earlier-described comparisons of wideband and narrowband circuits involved only a limited sample of equipments, the tests were made under carefully controlled conditions, and it is believed that the results are reasonably indicative of performance to be expected in the field. Those results suggest that a considerable price in performance has been paid for the additional channels obtained by channel splitting. As was shown, under fairly severe impulse noise interference and with equivalent modulation levels, the narrowband channel required 12 or 13 db less path loss than the wideband, to give a marginal circuit of comparable quality. This corresponds to a reduction in limiting coverage range of some 25% to 50%, which is a sizable loss. Of course the average effect will be less than this, and the mobile radio user who is fortunate enough to be in a quiet location may actually find that narrowband has improved the range; but that is cold comfort to the one who must operate against severe noise interference. It appears that any consideration of proposals for further channel splitting in the mobile bands should include a very thorough evaluation of impulse noise effects.

Acknowledgements

Most of the work described in this paper was done by Messrs. R. V. Crawford, V. A. Douglas and H. J. Bergmann of Bell Telephone Laboratories, to whom this acknowledgement is gratefully extended.

---

* That is, noise with a peak voltage 60 db above 1 microvolt, per megacycle of bandwidth.

** Quasi-peak measurements of impulse noise voltage are obtained by adjustment of the time constants in the noise detector, to give weighted values which will fall somewhere between the true peak and the RMS voltages.

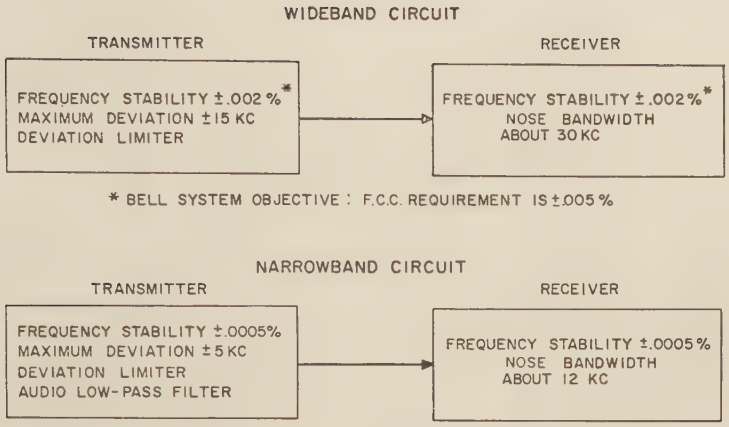


Fig. 1 - Main Features of Wideband and Narrowband Circuits

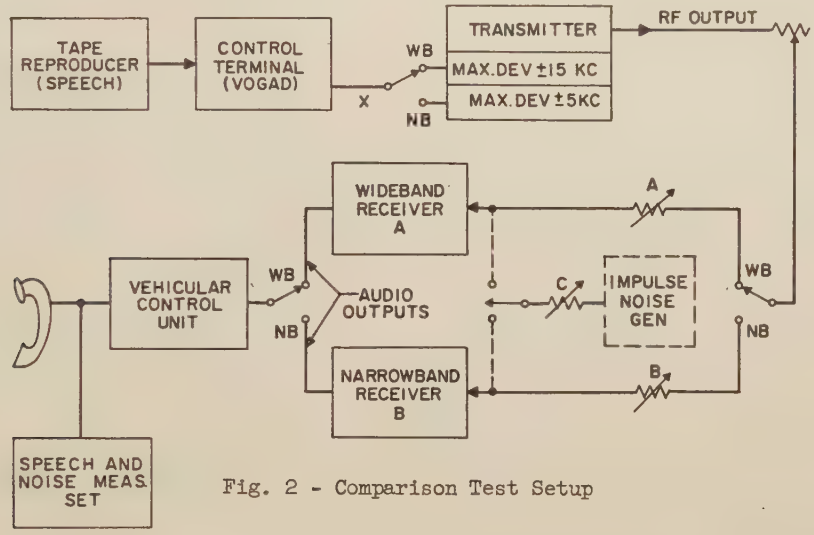


Fig. 2 - Comparison Test Setup

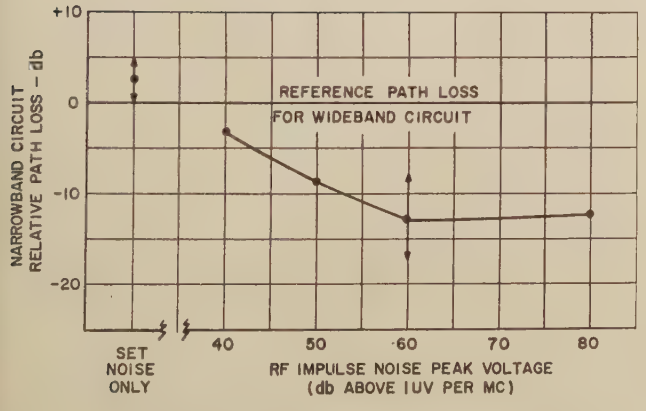


Fig. 3 - Comparative Performance of Wideband and Narrowband Circuits

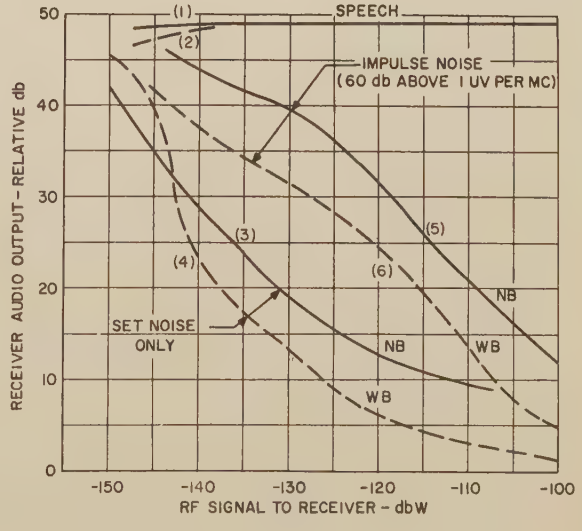


Fig. 4 - Receiver Characteristics

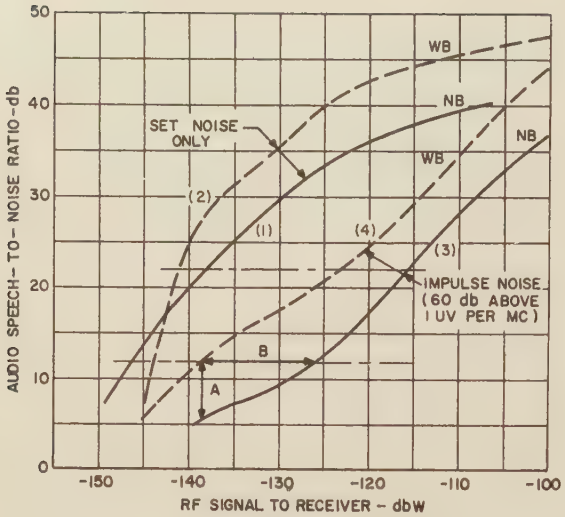


Fig. 5 - Speech-to-Noise Ratios

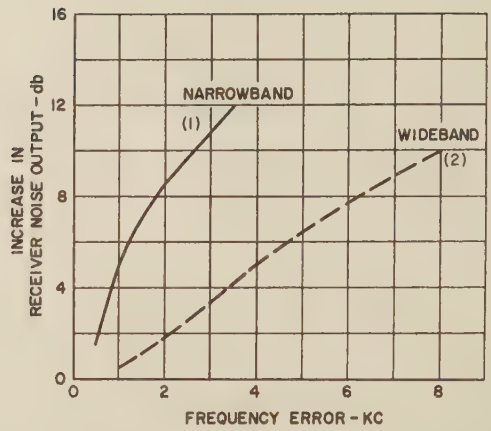


Fig. 6 - Noise Penalty of Frequency Errors

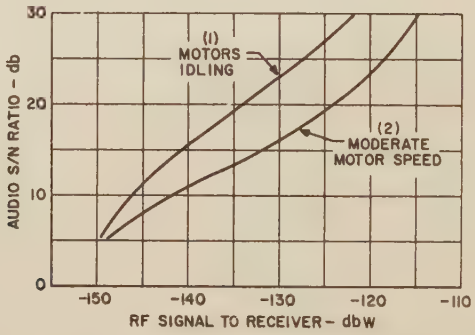


Fig. 7 - Variations in Effect of Constant-Peak-Voltage Impulse Noise

A TRANSMISSION LINE AND RADIATING SYSTEM MEASUREMENT

W. F. Biggerstaff
U. S. Forest Service
Beltsville, Maryland

Summary. One of the major problems in mobile system maintenance is adequate evaluation of the transmission line and antenna performance. This paper describes a practical method of technique for making line loss and antenna match measurements to assure proper installation and continued satisfactory performance. Certain innovations have been incorporated in the process which are calculated to provide the maximum information with a minimum effort.

The procedure consists essentially of making two transmission line input admittance measurements at widely differing frequencies. The results are plotted on a simplified version of the Smith Chart in a manner that yields transmission line loss and antenna match. These two properties adequately evaluate the transmission system for maintenance purposes.

Both measurements are made at the input end of the transmission line without altering or disturbing the antenna or load end of the line. This is important where tall towers are involved or the antenna is otherwise not readily accessible. Accuracies are more than adequate for service purposes and the results are consistent under a wide variety of conditions. A simple technique for determining the distance in feet to a fault such as a break or short circuit is also described.

Introduction

Defects, improper performance and incorrect adjustments in the basic radio transmitter and receiver of the usual mobile system are generally recognized without difficulty by the aid of techniques and instruments commonly understood. On the other hand, faults in the radio-frequency transmission line and antenna system are not always so easily recognized. As a result they frequently go undetected and the degraded performance is accepted as a normal condition. To bridge this gap in system maintenance, methods were investigated to more accurately evaluate the transmission line and radiating system performance.

The first instrument to yield consistent results under a wide variety of conditions was the Byrne bridge. This is a device that measures magnitude of impedance and phase angle at the point of attachment. These properties may then be plotted on the Z-Theta Chart with line loss and antenna match obtained in a manner similar to that used for the Smith Chart. Additional use disclosed that not only could line loss and antenna match be established but defective conditions could be identified, their position located, and corrective measures indicated in many instances.

The Byrne bridge was considered to be too bulky and costly for extensive use as a portable field instrument so other devices were investigated for such applications. The General Radio type 1602-B admittance meter was finally selected because of its smaller size, rugged construction, and moderate cost. The fact that it measures complex admittance instead of impedance should create no difficulty since a "Calculator" has been devised for the purpose of providing line loss and antenna match directly from quantities as read on the admittance meter.

The basis for this system of measurement stems from the fact that incident energy is reflected from any point in a radio-frequency transmission line at which at abrupt discontinuity in impedance exists. If such a line is terminated in an open circuit, short circuit, or a reactance of any value, the total incident energy at that point will be reflected. This condition is approached for most antennas when a measurement is made at a frequency that is substantially different from the system operating frequency. The result is a standing wave pattern which is dependent upon properties of both the transmission line and its antenna load. These properties may be determined by correct interpretation of appropriate meter readings.

Figure 1 shows the voltage standing wave pattern, impedance pattern on both rectangular and polar coordinates, and admittance pattern on polar coordinates for a radio-frequency transmission line having six decibels loss and terminated in an open circuit. A short circuit termination produces essentially the same pattern except that it will be displaced by one-quarter wavelength. It is apparent that the impedance (or admittance) variations along the line under the mismatched condition oscillate between high and low values at alternate quarter-wave points with decreasing amplitude when attenuation is involved. The result, when plotted on polar coordinates, is a logarithmic or exponential spiral with a radius inversely proportional to the attenuation. This provides a method for graphically determining line loss since a radial arm can be pivoted at the center of the chart and

calibrated directly in appropriate loss values.

Instruments

Accessories required for use with the admittance meter include a radio-frequency signal source, a detector, and the necessary cables, charts, connectors, etc. A complete instrument complement is shown in Figure 2 and Figure 3.

Signal Source. The radio-frequency signal source may be a standard signal generator having an output of at least one-tenth volt at an impedance of approximately 50 ohms. The generators recommended by the General Radio Company are the 1211 and 1215 unit oscillators. Another signal source found to be satisfactory, although somewhat low in power output, is the Measurements Corporation model 59 megacycle meter or grid-dip oscillator. A shielded enclosure, which incorporated the coupling loops, was constructed for the RF head. The coupling loops were resistance compensated to provide a 50-ohm source impedance for the connecting cables.

Detector. The detector is essentially a receiver having an output characteristic proportional to signal input in order that a null may be observed. It must be broad-band if other than crystal-controlled signal sources are used; it should have a sensitivity of better than 10 microvolts; it must be well shielded against leakage; and it must cover the measuring frequencies of interest. The Hewlett-Packard model 417-A VHF detector is a very good instrument for this purpose. The AN/APR-1 and AN/APR-4 search receivers are suitable if leakage is avoided.

"Calculator". The transmission line loss and antenna match calculator, shown in Figure 4, is a simplified version of the 20-millimho characteristic admittance Smith Chart. Since the sign of the reactive component is not required for line loss and antenna match measurements, one-half of the set of coordinates has been eliminated and the chart becomes a semicircle. A radial arm is pivoted at the center on which is inscribed the DB loss and SWR scales. SWR is used to indicate the degree of match and should not be confused with power standing wave ratio which is sometimes abbreviated SWR. Its use in this manner is based on the fact that the ratio of a resistive load and line characteristic impedance is numerically equal to the voltage standing wave ratio at that point. This has been abbreviated SWR and used as a measure of load match. A perfect match is indicated by a ratio of unity.

Measurements

In practice, it is convenient to make the transmission line loss measurement first. The process consists of adjusting the conductance and susceptance controls of the admittance meter for null with the generator and detector set to approximately one-half the system operating frequency. The values of conductance and susceptance are transferred to the appropriate coordinates on the "Calculator". The radial arm is then rotated to the plotted intersection and the transmission line loss read in decibels. Since the measuring frequency is one-half the operating frequency, a correction is necessary and the loss must be multiplied by 1.5 to obtain the loss at the operating frequency.

It is not always possible to obtain a satisfactory null at frequencies below 30 megacycles. Under such conditions it may be necessary to double the operating frequency and divide the loss by 1.5 to obtain the loss at the operating frequency.

Antenna Match. The antenna match measurement is made in exactly the same manner as the line loss measurement except that it is made at the operating frequency of the antenna. The values of conductance and susceptance, as read, are transferred to the "Calculator" and the degree of match noted on the SWR scale. If an appreciable loss exists in the transmission line, the SWR reading must be increased by an amount proportional to the DB loss. Since the two scales are side by side, this is easily accomplished.

Other Measurements. There are many defects and faults in the transmission system that may be readily recognized as the operator becomes proficient in the use of the instruments. For instance, if the match improves as the measuring frequency is reduced below the operating frequency, the antenna is incorrectly adjusted (probably too long). A transmission line of incorrect characteristic impedance will be indicated by plotting as a circle on the conventional Smith Chart for frequencies above and below the normal operating frequency but the center of the circle will be in the vicinity of the characteristic admittance of the line being measured.

Short sections of line, pads, adaptors, connectors, etc. may be checked by terminating them in a resistance equal to their characteristic impedance and measuring the input admittance. Lines should be flexed during the check to avoid overlooking an intermittent condition. An intermittent condition in an antenna may be checked from the ground by making a line input admittance measurement during windy weather.

Faults that cannot be readily recognized and other anomalous conditions may be analyzed by making a series of measurements at small frequency increments above and below the normal operating frequency of the antenna. The conductance and susceptance values are then plotted on the standard Smith Chart and the

points connected by a smooth curve. Any abrupt discontinuity in the curve indicates an error or insufficient plotted points.

Characteristic Impedance. The characteristic impedance of an unknown transmission line may be checked by varying the measuring frequency sufficient to make an input admittance measurement at an adjacent node and antinode. These are identified as zero susceptance points. The characteristic admittance will be the geometric mean of the two values of conductance and the characteristic impedance will be the reciprocal of admittance. The line should be long enough to have at least three decibels loss at the measuring frequency or else terminated in a resistive load to avoid extreme values of conductance.

Line Length. The frequency increments necessary to plot a smooth curve on the standard Smith Chart must be smaller for long transmission lines than for short ones. Investigation revealed that there is a mathematical relationship between the frequency difference required to measure two adjacent nodes or antinodes on a mismatched line and the line length. This forms the basis for a convenient method to measure the distance to an abrupt discontinuity in the line such as a short circuit or a break. The velocity factor of the line must be known with reasonable accuracy.

This method consists of adjusting the frequency of the generator and detector until a null is obtained with the susceptance control set to zero and the conductance control to a low value. The frequency should be noted and then increased slowly until the next null is obtained. The frequency separation between the two nulls is divided into a quantity consisting of 492 times the velocity factor of the line to obtain the distance in feet. For solid dielectric lines such as the RG-8/U and RG-11/U having a velocity factor of 65.9%, this becomes 324 divided by the frequency difference. The best measuring frequency will vary somewhat with line length but anything from 50 to 200 megacycles is usually satisfactory. Greater accuracy may be achieved and errors avoided if a series of points are plotted at smaller frequency increments on a standard Smith Chart and connected by a smooth curve. The frequency change necessary to complete a circle on the curve will indicate the frequency separation between the two nulls. The greatest accuracy is achieved with an open circuit or short circuit termination.

Other Impedance Systems. So far, all measurements have been based on a 20-millimho characteristic admittance (50-ohm impedance) system. The General Radio Company recommended that a half-wave section of line be used with the admittance meter when other than 50-ohm systems are involved. Such a section effectively transfers the measuring point within the instrument to the point at which it connects with the "unknown". This has not been found necessary for 75-ohm systems (13.3-millimho) and, if used, complicates the measuring procedure considerably since the line must be changed or adjusted for each change in frequency.

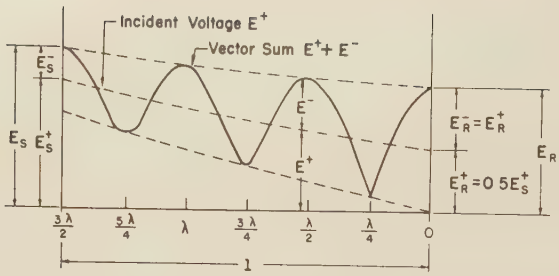
The 20-millimho "Calculator" may be used to obtain line loss and antenna match in a 75-ohm system if each value of conductance and susceptance is first multiplied by 1.5. This effectively "normalizes" the readings to a 20-millimho reference. All other measurements for a 75-ohm system should be plotted on a normalized Smith Chart after first dividing each conductance and susceptance reading by 13.3. This may be reduced to a very simple slide-rule operation by multiplying each reading by 0.075 before plotting.

Conclusion

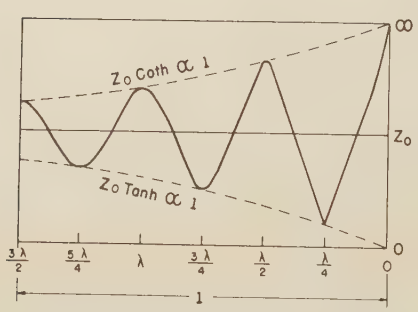
Accuracies of this system of measurement are, in general, more than adequate for radiation system analysis and service maintenance work. Instrument costs, while some might consider high, should not price the method out of service organizations with antenna maintenance responsibilities. The flexibility in the choice of accessories, the multiple use nature of some of them, and the results that can be achieved in evaluating a portion of the overall radio system that has, in many instances, remained heretofore unknown, can more than repay the investment. It is felt that this paper describes only a few of the many possibilities for use of these measuring methods and, as experience is gained, other useful data will be found.

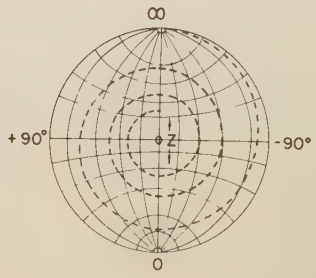
References

1. General Radio Company, Operating Instructions Type 1602-B Admittance Meter.
2. Electronics, Vol. 12, January 1939.
3. Electronics, Vol. 17, January 1944.
4. Electronic Measurements, Terman & Pettit.
5. Circuit Analysis of Transmission Lines, Stewart.
6. Electronic Designer's Handbook, Landee, Davis, & Albrecht.



Voltage, Impedance, and Admittance variations along a radio-frequency transmission line of length 1 having 6 DB loss and terminated in an open circuit ($Z_L = \infty$). Loss per wave-length magnified.





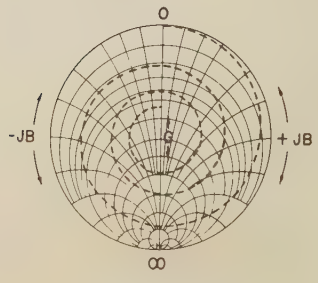


Fig. 1 - Mismatched Transmission Line Patterns

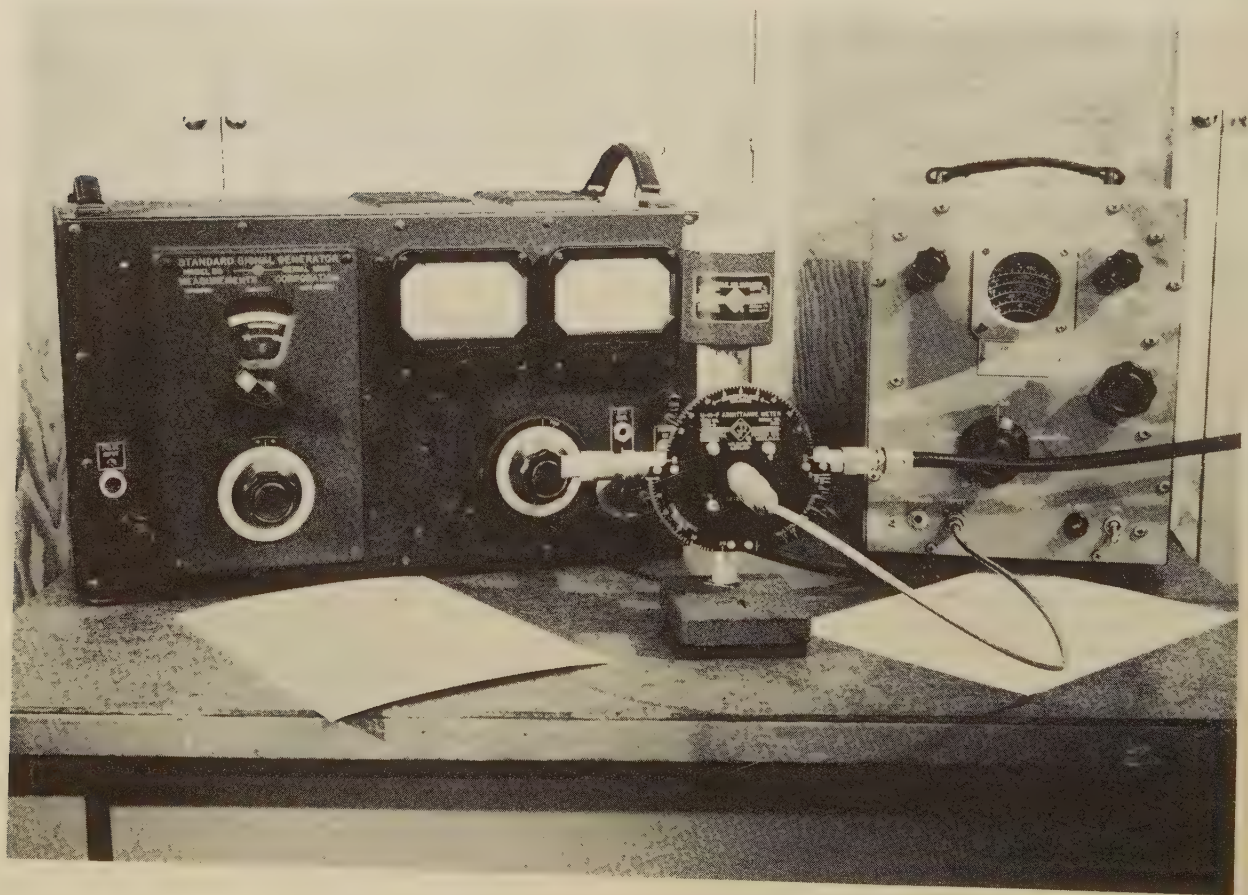


Fig. 2 - Setup Using Standard Signal Generator

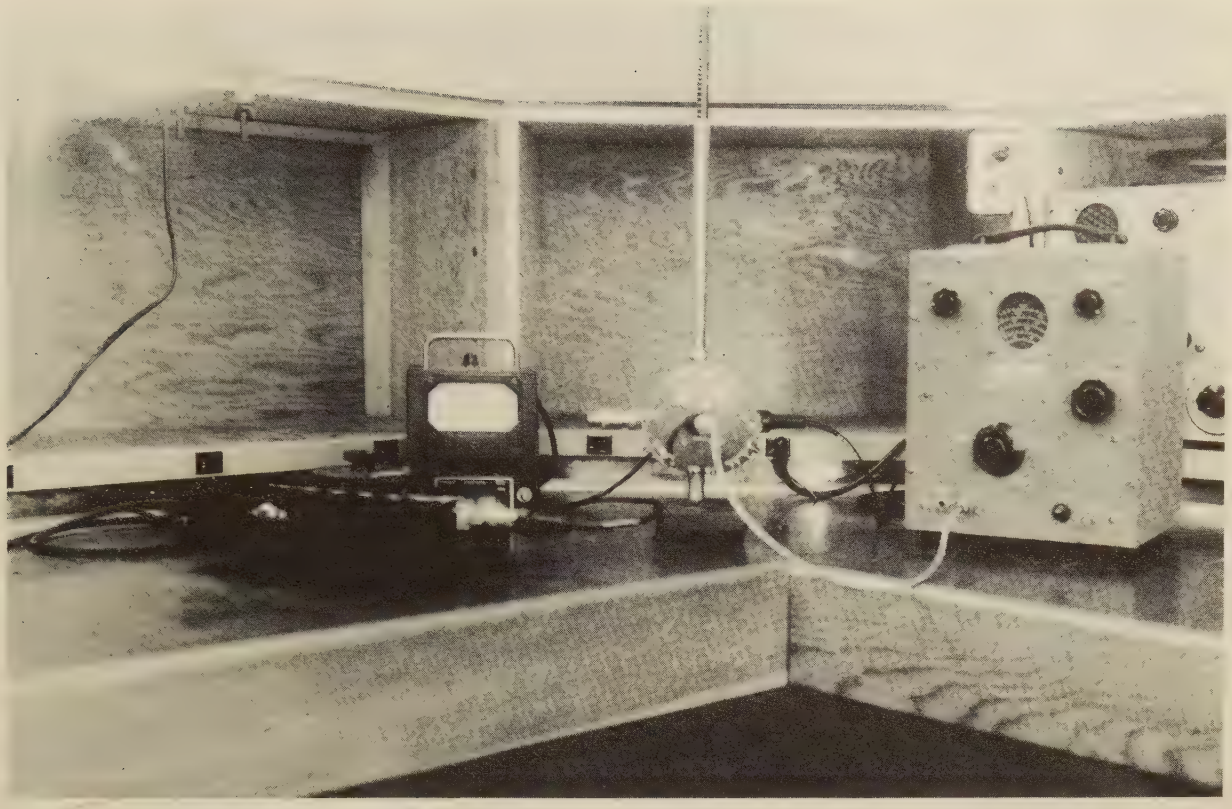
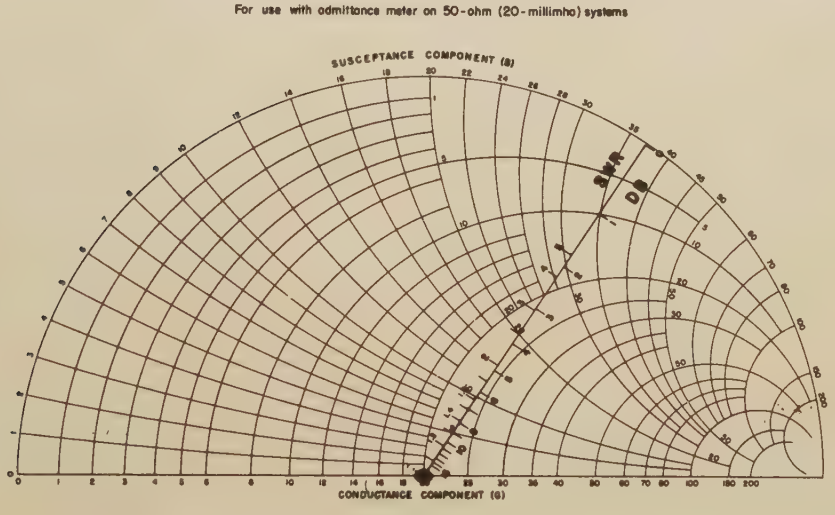


Fig. 3 - Setup Using Grid Dip Oscillator



INSTRUCTIONS

1. Connect admittance meter and accessories.
2. Set signal source and detector to 1/2 antenna frequency.
3. Balance admittance meter and plot values of G and B on chart.
4. Rotate radial arm and read DB loss at plotted point.
5. Multiply loss by 1.5 to obtain loss at operating frequency.
6. Repeat step 3 at the operating frequency and plot on chart.
7. Rotate radial arm to plotted point and note DB scale reading.
8. Increase the distance from the center on the radial DB scale by the number of DB found in step 5 and on the opposite scale read the corrected SWR.

Fig. 4 - Transmission Line Loss and Antenna Match Calculator

AN OMNIDIRECTIONAL MEDIUM GAIN 460 MC BASE STATION ANTENNA USING SIMPLIFIED SUPPRESSORS

J. S. Brown
Andrew Corporation
Chicago, Illinois

Summary

The use of suppressors to provide a collinear array that has a single feed point has been described[1]. The suppressor design reported was one that provided optimum performance. During the course of the development program, several other types of suppressors were investigated. One type, although slightly inferior in performance to that described, is very attractive because of its simplicity. While it is not satisfactory for use in high gain (10 db) antennas, it is suitable for use in medium gain (4-6 db) antennas, and offers interesting cost reduction possibilities. The design and performance of a 6 db antenna for use in the 450-470 MC band is described.

A suppressor is essentially a device, located on a long conductor, that breaks up the conductor into a series of half-wave in-phase radiators. It suppresses the radiation currents that would flow on the long conductor which would be 180 degrees out of phase with the desired currents. If a single feed point array is to be realized, the suppressor must also provide the feed system by coupling between the ends of adjacent radiators along the array. The suppressor previously described is shown in Figure 1. The two quarter-wavelength chokes on the outside provide good suppression of out-of-phase currents, and the coaxial feed system composed of the inner rod and the cylinder immediately adjacent to it, provides a very effective system for end feeding from one half-wave radiator to the next. However, its disadvantages are that it is relatively costly to manufacture and is also a half wavelength long. A shorter suppressor would result in a shorter overall antenna array.

A simplified suppressor was developed, and it appears in Figure 2 in schematic form[2]. It consists of a dielectric loaded length of coaxial transmission line. The ends of the inner conductor are connected to the ends of adjacent radiators. The suppressor is a quarter wavelength long physically. The space between the inner and outer conductor is filled with a dielectric having a dielectric constant of 4, so that it is a half wavelength long electrically. This provides in-phase feeding of adjacent radiating elements. One of the commercially available Fiberglas and resin compositions meets this requirement for dielectric constant.

Since the outer conductor is only a quarter wavelength long physically, it tends to suppress radiation currents from flowing on its outer surface. The inside surface of the outer conductor and the outside of the inner conductor form a coaxial feed line that provides the neces-

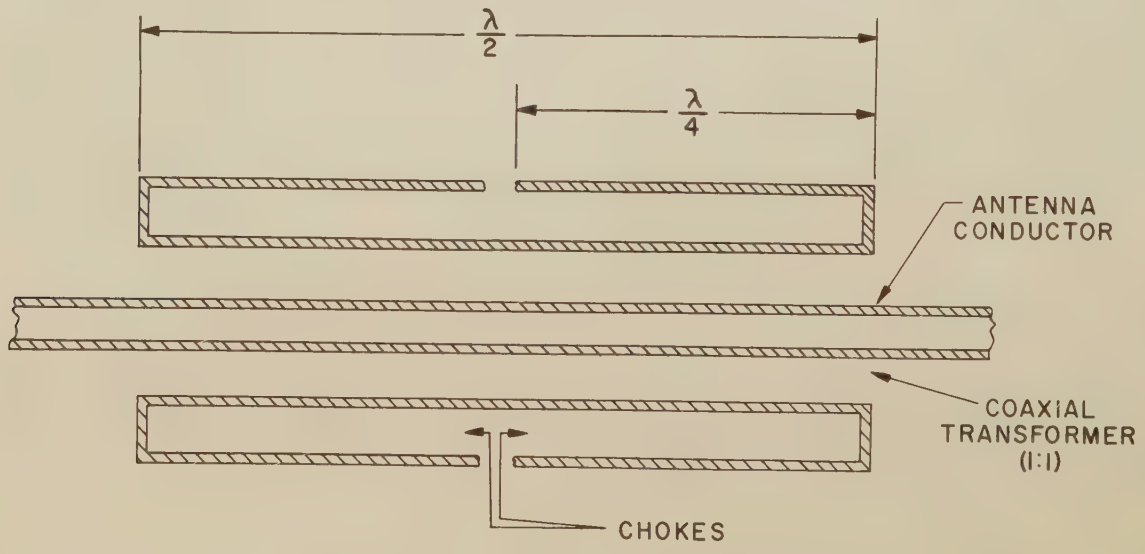


Fig. 1 - Optimum Suppressor Cross Section

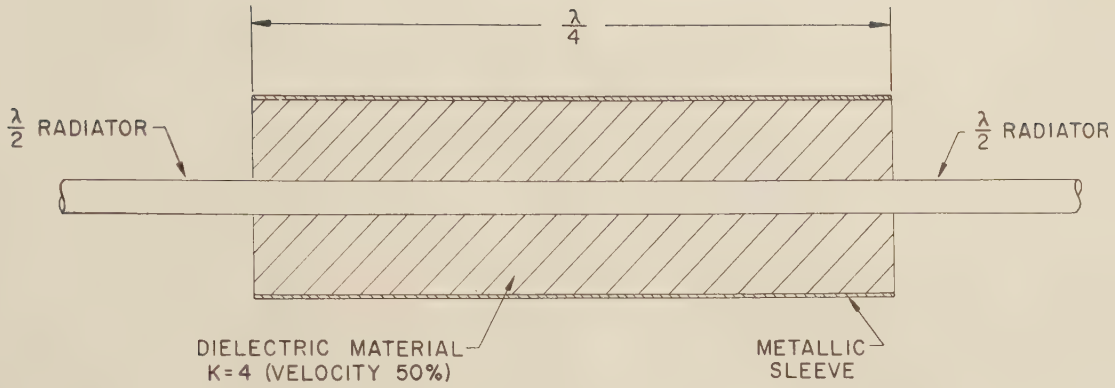


Fig. 2 - Simplified Suppressor

sary phase delay for feeding energy from one radiator to the next in correct phase relation. It also provides adequate coupling, so that reasonable efficiency of the aperture is realized for gains up to about 6 db. For higher gain (longer) arrays, the illumination taper at the end is so great that the efficiency of the aperture is reduced to impractical values.

It can be seen that the performance of this suppressor is not as good as the unit described previously, by comparison of the radiation patterns shown in Figures 3 and 4. Figure 3 is the pattern of an array using three radiating elements and two simplified suppressors, while that of Figure 4 is for the same array using the more elaborate suppressors. It can be seen that the beam width of the array using the simplified suppressors is broader, indicating less efficient coupling from the center radiator to the end elements. The side lobe structure can also be seen to be less clearly defined, which indicates that some out-of-phase currents are flowing on the array. However, the performance is adequate for the antenna to be described below.

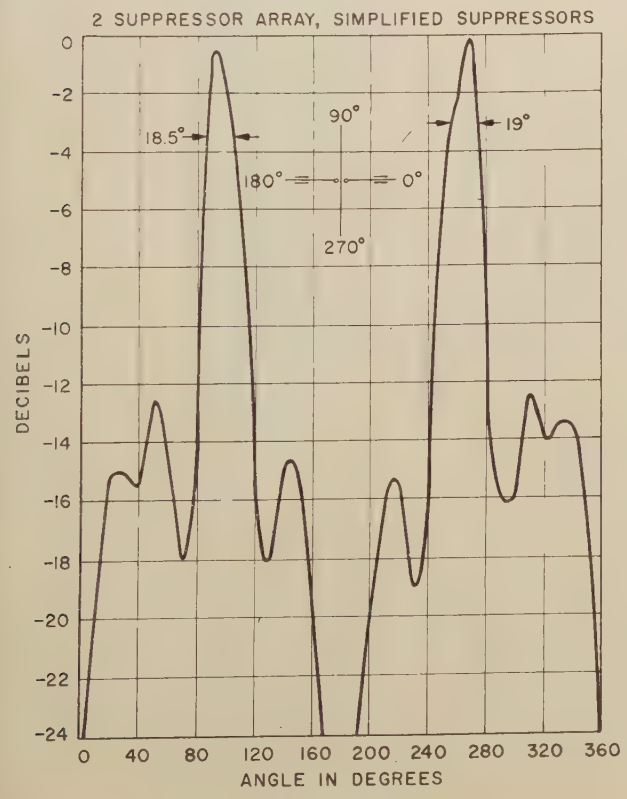


Fig. 3 - E-plane Pattern (Simplified Suppressors)

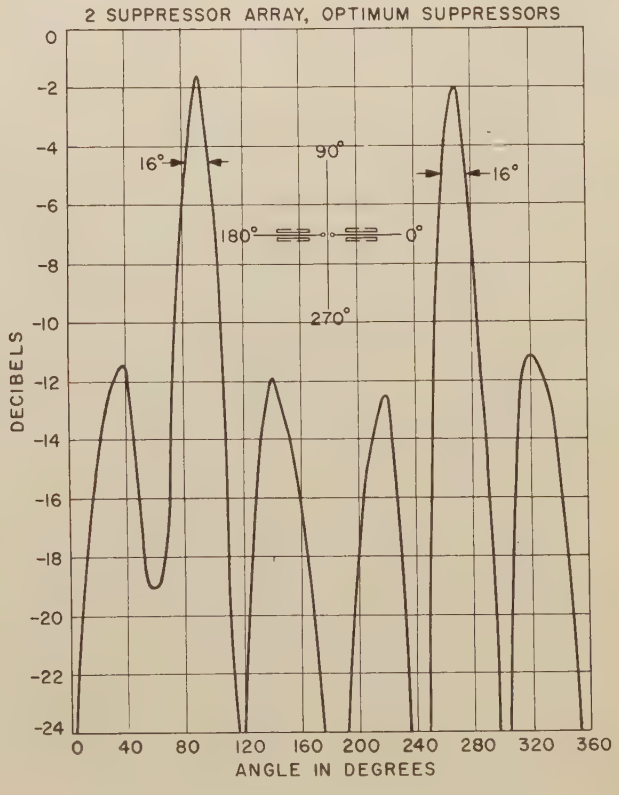


Fig. 4 - E-plane Pattern (Optimum Suppressors)

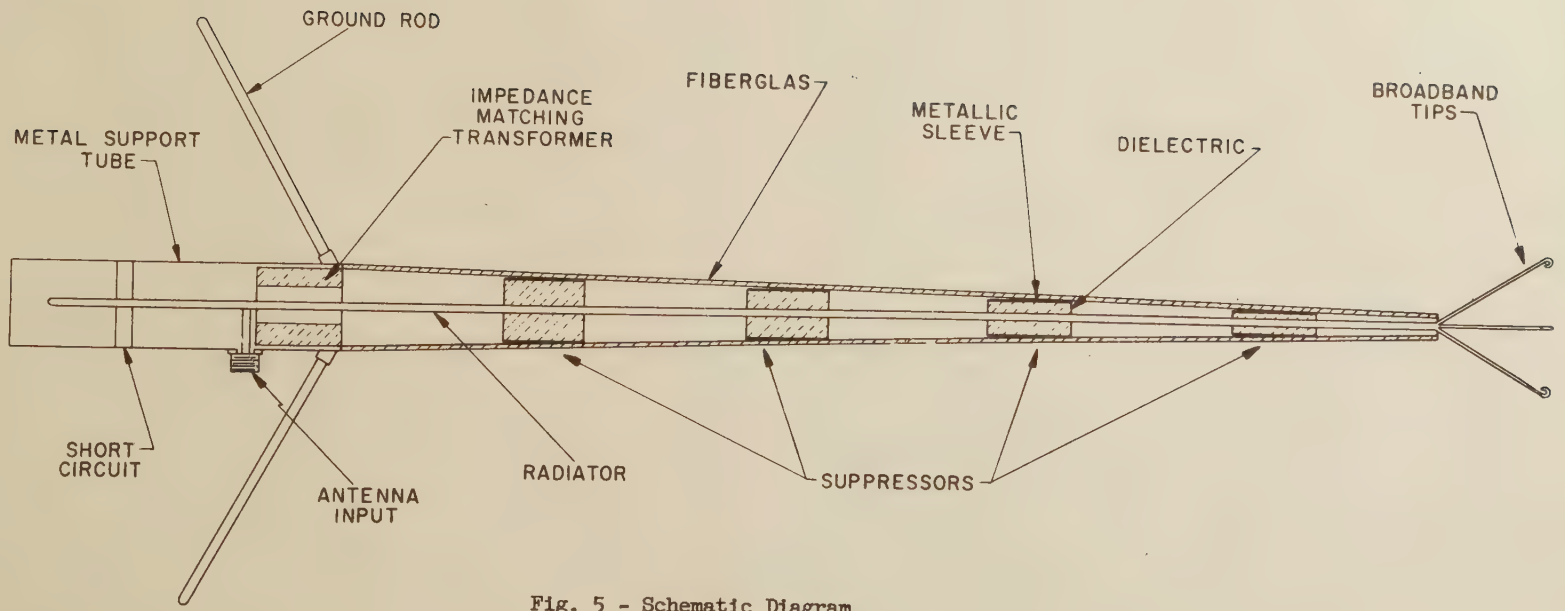


Fig. 5 - Schematic Diagram

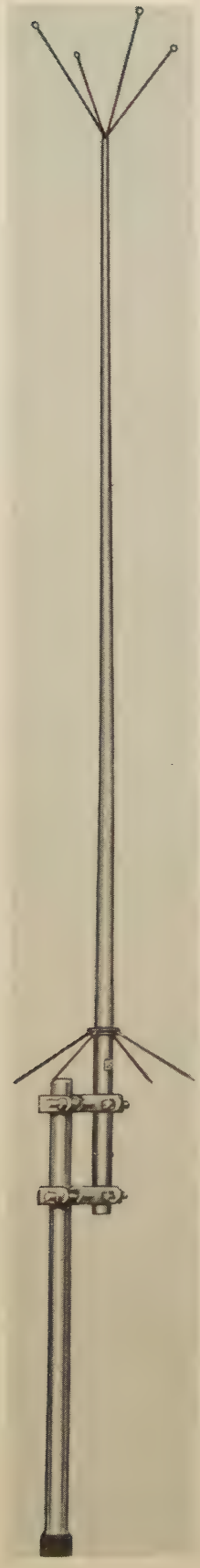


Fig. 6 - Type 206 Antenna

A commercial design for 450-470 MC, utilizing this principle, has been developed, and it is shown in the schematic diagram of Figure 5 and the photographs of Figures 6 and 7. It consists of five half-wave radiators and four suppressors. The array is end fed against a ground plane. A quarter-wave shorted stub extends below the ground plane and provides not only impedance matching, but also a dc ground for lightning protection. The input connection, shown in Figure 7, is a shunt fed arrangement tapped in at the proper place on the shorted stub. The stub also provides a convenient location for an impedance matching transformer which is shown in the schematic diagram. Several other impedance matching devices are incorporated in the final design, but are not shown in the schematic diagram.

It can be seen from the photograph of Figure 6, that both the top radiator and the ground plane are conical in configuration, being formed by sloping rods or wires. This construction provides two advantages. Since a conical radiator is inherently a broadband configuration, the impedance bandwidth of the array is improved considerably. It was also found that the gain of the array was increased significantly. The reasons for this appear to be two in number; one, the aperture is increased because of the sloping ground plane, and two, the top radiating element apparently provides better loading of the radiators below it, and the result is a more uniform illumination of the entire aperture.

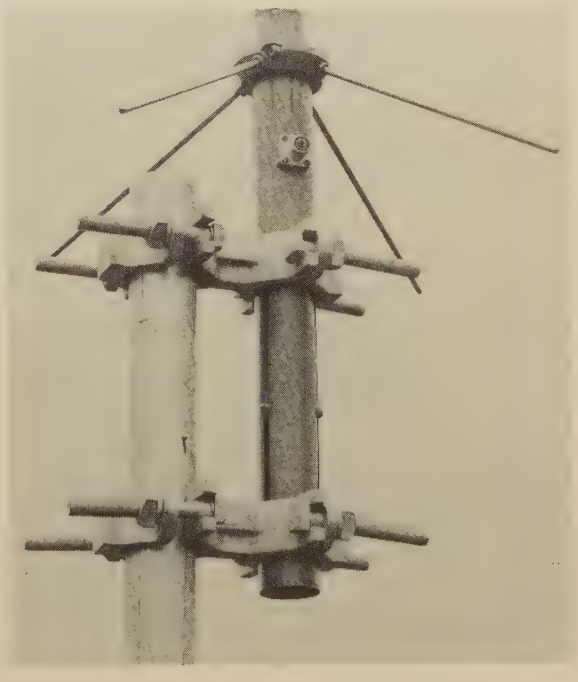


Fig. 7 - Base and Input

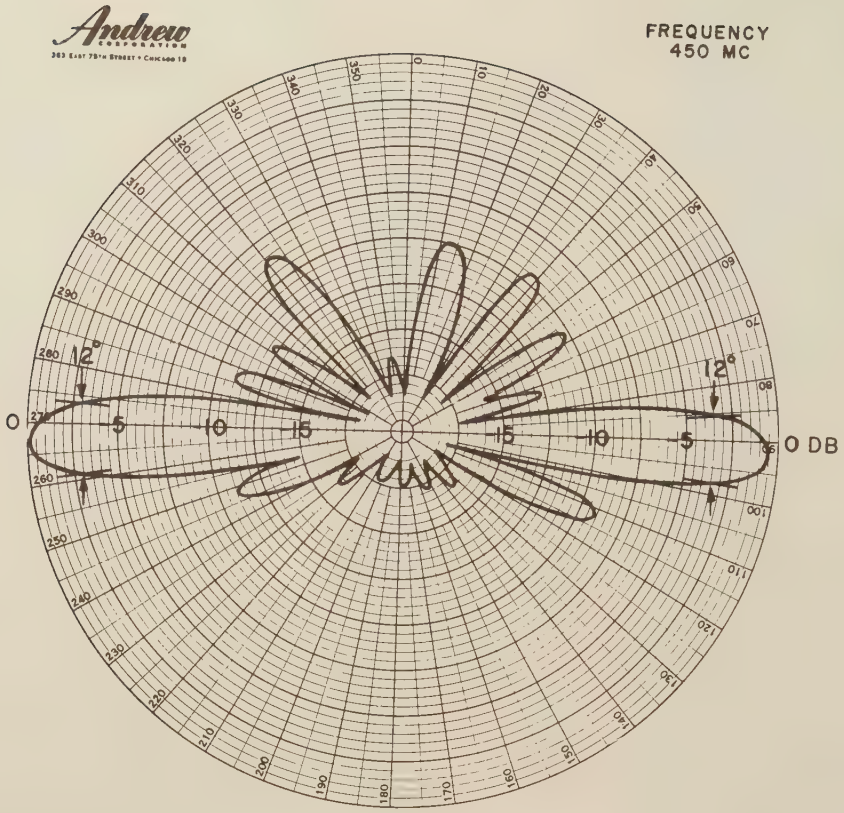


Fig. 8 - Vertical Pattern

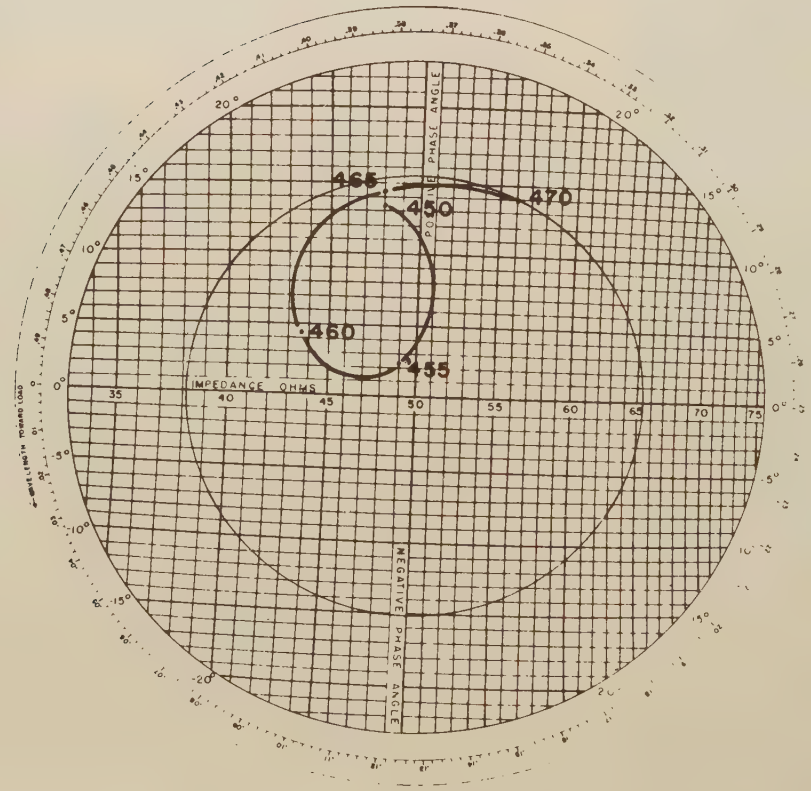


Fig. 9 - Imput Impedance

Structurally, the outer Fiberglas tube provides the entire mechanical support of the antenna. The inner rod is quite weak, being only 1/4 inch in diameter. The suppressors are solid Fiberglas and the outer conducting sleeves of the suppressor are made of perforated foil. In the region of the radiators, air spaces are provided between the radiators and the outer Fiberglas tube in order to keep weight, dielectric losses, and cost to a minimum. A system of drain tubes is provided from the top to the bottom of the array, so that no moisture can accumulate in the air spaces.

The total weight of the antenna is 11-1/2 pounds. The antenna utilizes brass, copper, and stainless steel on all exposed metallic surfaces to provide a maximum of corrosion resistance in salt spray environments. The over-all length of the antenna is 112 inches, and the height above the ground plane is 96 inches. The antenna is designed for a windloading of 30 pounds per square foot with 1/2 inch of radial ice.

The average gain of the array is better than 6 db over a half-wave dipole across the 450-470 MC band. Typical radiation pattern and VSWR measurements on a production unit are shown in Figures 8 and 9. Since the array is end fed, beam tilt across the frequency band would be expected. The curve shown in Figure 10 gives the theoretical and actual measured beam tilt. It can be seen that this is not a disadvantage for an array with this gain.

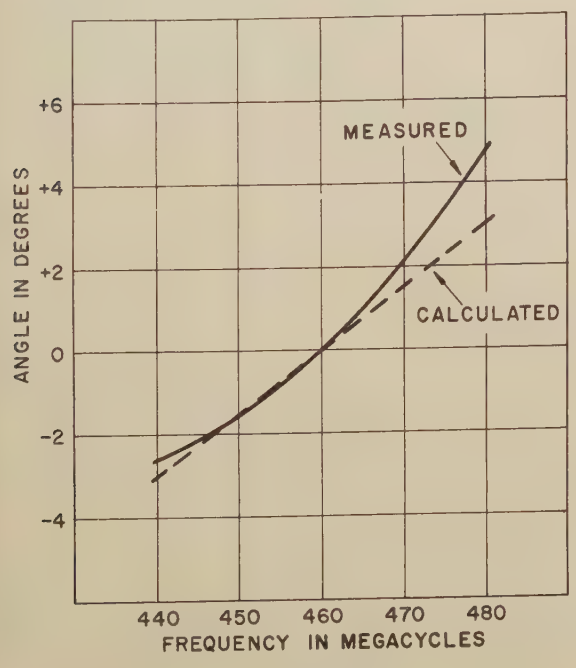


Fig. 10 - Beam Tilt

The Fiberglas construction of the suppressors employs relatively high loss insulating material. Therefore, determination of gain by radiation pattern integration alone is not adequate, since this calculation does not take into account ohmic losses within the antenna structure. It was therefore necessary to determine gain by a comparison measurement.

The method used was to compare the signal received from this antenna with the signal received from an antenna that was known to be essentially lossless. The gain of the lossless antenna was determined by pattern integration, and the difference in the two signals, with the appropriate calculated corrections for relative power input, provided a gain measurement for the array under consideration. Comparison of this figure with the gains obtained from pattern integration provide a measure of the loss due to dielectric and ohmic losses in the antenna itself. This loss appears to be approximately 1/4 db. The rated gain of 6 db includes allowance for this loss.

The gain of the final design was influenced by several factors. The theoretical gain of five half-wave radiators is almost 8 db for one wavelength center-to-center spacing. The gain of the array was reduced somewhat from this figure, since it was necessary to space the radiators closer than one wavelength in order to use the feed system desired. The ohmic loss described above is also a contributing factor. The conical top radiator and ground plane have the effect of increasing the gain, for the reasons described above. The ultimate determination of gain, therefore, must be made experimentally.

Conclusion

The simplified suppressor provides a satisfactory design approach for a simple single feed point low cost medium gain antenna.

Acknowledgement

Acknowledgement is made to M. W. Scheldorf, who conceived the original design concept and supervised the development work of the antenna design.

References

1. M. W. Scheldorf "New Type High Gain Station Antenna" IRE Transactions on Vehicular Communications, PGVC-11, July, 1958, pp10-19.

2. Patent 2,896,206, M. W. Scheldorf, July 21, 1959. Other patents pending.

STRICTLY PERSONAL

George M. Dewire
General Electric Company
Lynchburg, Virginia

Summary

Developments in miniaturization in radio design have been aimed for years at personalizing the equipment. The design parameters are affected by not only technical accuracy but also somewhat offset by personal likes and dislikes of the user. This paper describes a personal receiver in terms of development to achieve personal compatibility.

Introduction

The title refers to radio ... personal radio. The term "personal" here means "carried on-the-person". It does not mean carried by the person, it means worn. The achievement of the effective size for wearing is in itself cause for wonder in spite of the fact that it was predictable. The weight and size problem blocked all efforts until transistors reduced the battery complement, thus both size and weight. Today's limitations are still batteries to some extent but the size and weight of packaging have become a real factor. The next development phase will probably be further refinement and miniaturization in the device itself plus systems tailored especially to fully utilize it effectively.

The applications of personal radio are all about us, and the challenge of increasing the economic productivity of this nation is linked directly to the efficiency of communications. We are today witnessing the fruits of vehicular communications. In almost every business and industry personal communications will be that much more beneficial.

Back in 1957 in Syracuse the use of loupe and tweezers resulted in the assembly of a miniature receiver. Since then the basic approach has been refined and a product evolved which will soon be on the market. In the process of this development certain facts have come to light peculiar to personalizing radio, which have a direct effect on the design parameters from a technical standpoint. As is the pattern of communication development, the receiver comes first and it will serve the purpose very well to analyze one for "personal" peculiarities.

Audio Output

The business end of any radio is the end that talks. But how it talks is varied. Generally, there are three areas of listening requirements: private, semi-private and non-private under quiet, noisy, and very noisy conditions.

Personal radio is intended to be mostly private. Technically, the best reproduction and the best efficiency are obtained by an earpiece, inserted in the ear. Operationally, this may have adverse effects due to the interference with normal hearing. In any case, people are not enthusiastic about wearing something in their ear, so that what is best is not generally acceptable. Next best is the placement of the earpiece just outside the ear by a clip. This yields privacy, does not affect the ambient noise appreciably and is much more comfortable. This approach is not enthusiastically received either, but offers the best alternative in many cases, particularly in noisy areas.

For those whose application is semi-private or non-private the horn attachment to the earpiece is the arrangement which performs best. The horn is easily worn and easily directed at the wearer's head. The directional character of the horn serves to boost the audio efficiency and also make it nearly impossible to understand from any other direction. In addition, it was found that the earpiece/horn outperformed any other device except the earpiece in or on the ear in actual high ambient noise level tests.

Antenna

The arrangement of antennas on personal equipment is closely related to the audio problem. To be a personal item, the receiver cannot be large. If it is small it certainly doesn't make much of an antenna of itself. It was found that people, in general, do not want to wear a projecting antenna, although this would provide the best operation. So the next best solution had to be examined. In this case it was found by exhaustive field testing that a loopstick of ferrite would perform admirably. The presence of the earpiece cord, however, when added as a broadly tuned circuit across the loopstick improves the performance significantly without adding to the basic package. Here again the earpiece yields the best results since pulling the earpiece cord up to the ear yields a nearly omni-directional field pattern. An important feature of this arrangement is that no matter where the cord is placed it will complement the loopstick and even wadded up does not degrade it.

Physical Description

The receiver itself is, of course, pocket-sized. (See Fig. 1). It includes the two most important controls - Volume and Squelch - right on the top, with the On-Off control ganged to the Volume potentiometer. The center jack delivers audio to the earpiece.

There is a wide clip whose gripping edge is a strip of cerrated rubber designed to hold to but not to injure fabric. While the unit may normally be worn on the belt there is a preference on the part of some users for pocket carrying and shirts are easily torn or worn through by metallic gripping schemes.

The box itself is constructed of high impact plastic (Cycolac) and its single securing screw may be sealed for tamperproofing. The battery access is on the side for easy replacement without exposing the basic works of the receiver.

Inside the unit are the basic elements (See Fig. 2) the antenna at the bottom, battery and polarized battery holder next, the receiver container, space for selective signalling, and the audio squelch circuitry. The three tight-fitting covers at right seal the receiver modules and minimize interstage coupling. The modules themselves are not soldered in and may be withdrawn (within the limits of the wiring) for examination of parts and for detailed measurements. Tuning and metering points are accessible through holes in the module covers.

Block Circuit

Referring to the block diagram (See Fig. 3) the unit is a full, double conversion, superheterodyne VHF-FM receiver. The ferrite antenna feeds directly into a double-tuned antenna transformer then into a GE 4JX3C509 tetrode mixer. Here the signal mixes with that from the 2N501 first oscillator to produce 8.7 mc, the high IF, and provide gain. A 2N384 is used as high IF amplifier with a double tuned transformer ahead of it and a single tuned transformer coupling into the second mixer, a 2N1107. After achieving selectivity in a four coil low IF filter, two stages of 290 kc low IF gain precede the two limiters (all 2N1107's) into the discriminator transformer. This is the extent of the receiver proper. The degree of the selectivity is determined by the transformers in the center board. Either wide or narrow band and up to 6 coil variations are possible. At this point, a three-way choice may be made. The addition of an audio board will provide normal audio paging, or a tone network may be added for selective signalling, or, both may be added for selective paging. Fig. 4 shows a signalling receiver with reset button while Fig. 5 shows the internal circuitry. This makes this unit a truly versatile device which will provide optimum flexibility and use in the various system applications in use.

Service

Throughout the receiver design the human element has played a major role. But in addition, contributions to the art were pioneered in retaining serviceability while achieving miniaturization. Now, to be sure, these units are not to be serviced in the more or less brute force method of tube servicing, but it will be quite easy for a person skilled in transistor work to maintain them. First of all, the majority of circuitry is available by removal of the metal shields. The boards may be removed and repaired, or disconnected and replaced. No unsoldering is required for removal, just care. For those situations which occur quite frequently where the customer's frequency is changed, note that the front end crystal plugs in, thereby simplifying frequency change as well as providing an aid to trouble-shooting.

Conclusion

All of these items are part of proper design consideration arrived at by careful analysis of the application. In this area where people are the "vehicles" the installation is more critical than ever before. Because people are not built for radios, radios must be built strictly personal.

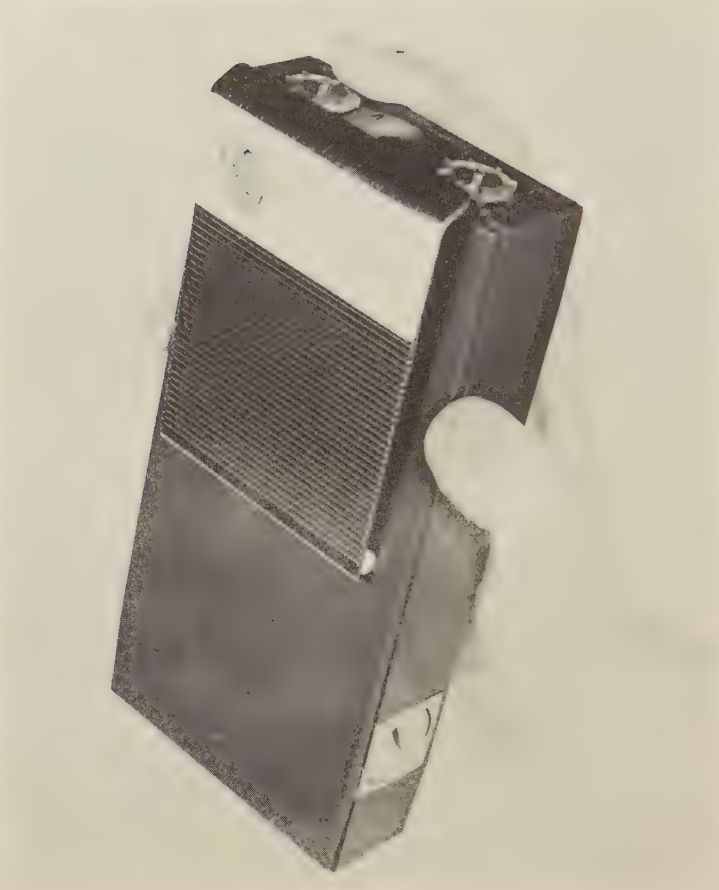


Fig. 1 - 150 Mc Personal FM Receiver - Exterior

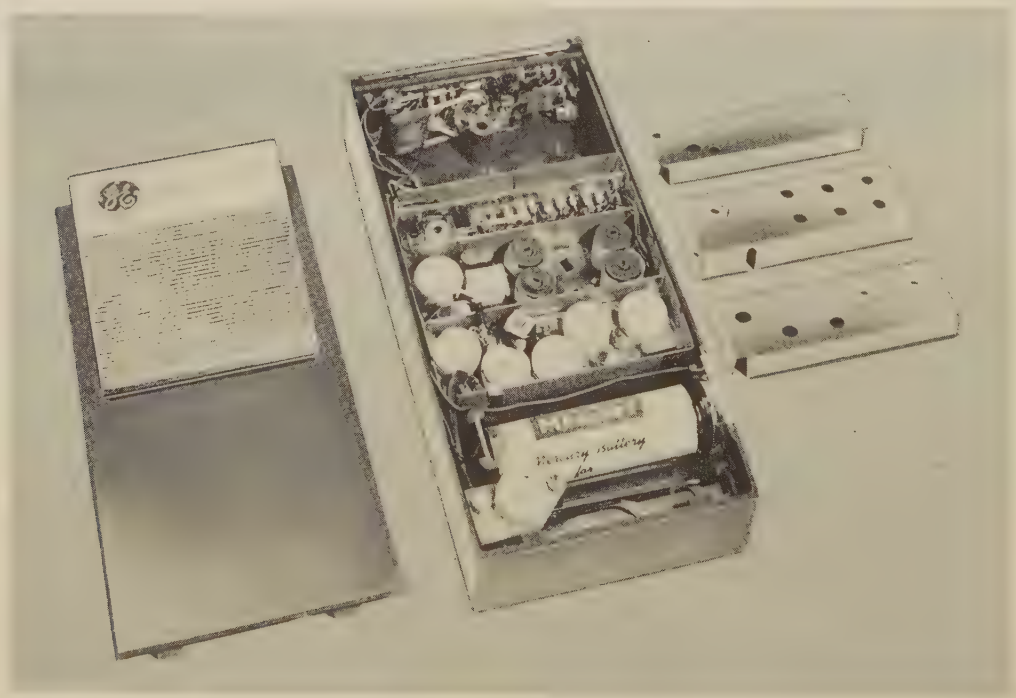


Fig. 2 - 150 Mc Personal FM Receiver - Interior

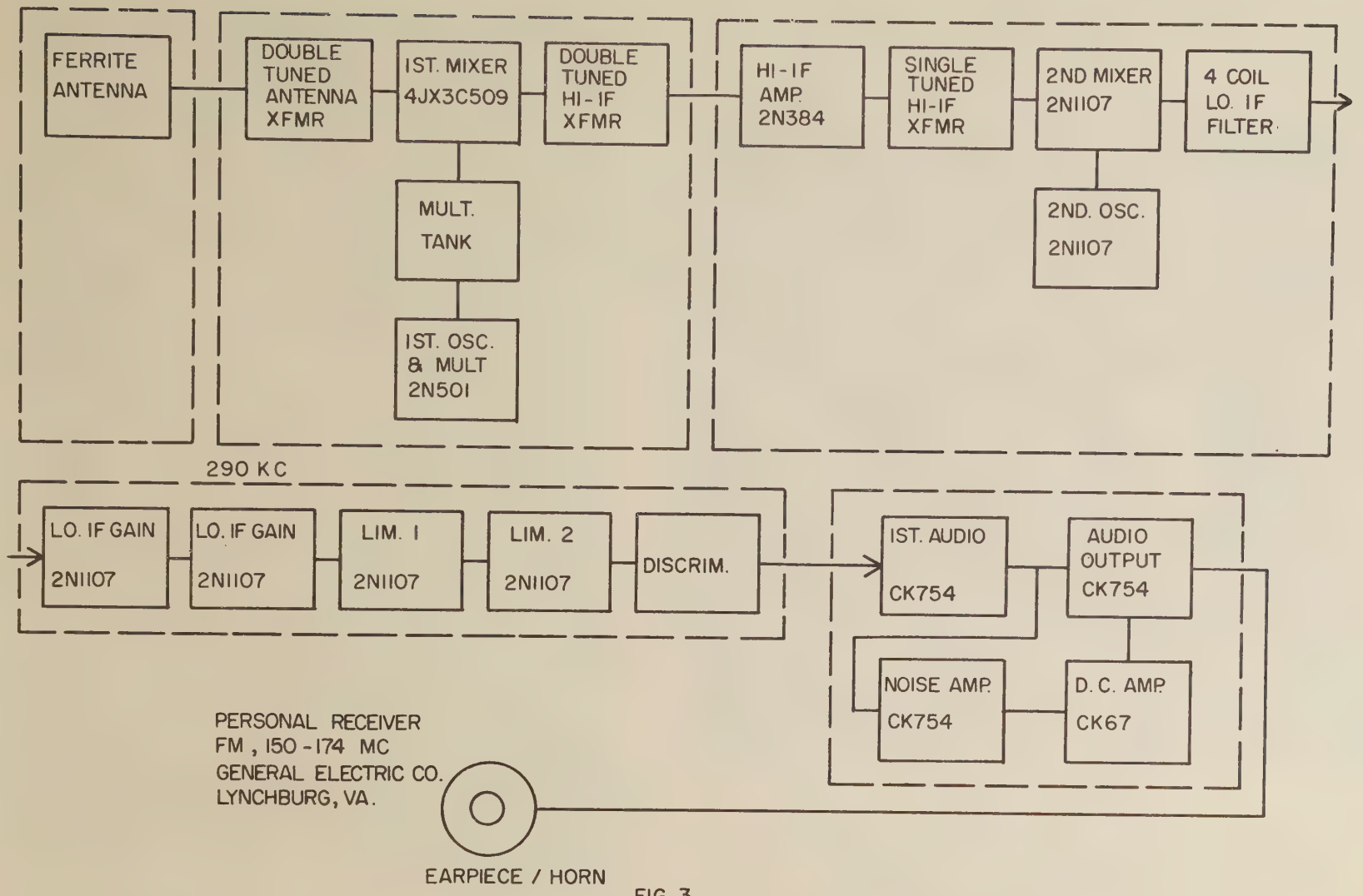

FERRITE ANTENNA
DOUBLE TUNED ANTENNA XFMR
IST. MIXER 4JX3C509
DOUBLE TUNED HI-IF XFMR
MULT. TANK
IST. OSC. & MULT 2N501
HI-IF AMP. 2N384
SINGLE TUNED HI-IF XFMR
2ND MIXER 2N1107
2ND. OSC. 2N1107
4 COIL LO. IF FILTER
290 KC
LO. IF GAIN 2N1107
LO. IF GAIN 2N1107
LIM. 1 2N1107
LIM. 2 2N1107
DISCRIM.
IST. AUDIO CK754
AUDIO OUTPUT CK754
NOISE AMP. CK754
D.C. AMP. CK67
PERSONAL RECEIVER
FM, 150-174 MC
GENERAL ELECTRIC CO.
LYNCHBURG, VA.
EARPIECE / HORN

FIG. 3

Fig. 4 - 150 Mc Personal FM Receiver with Signalling and Reset Button

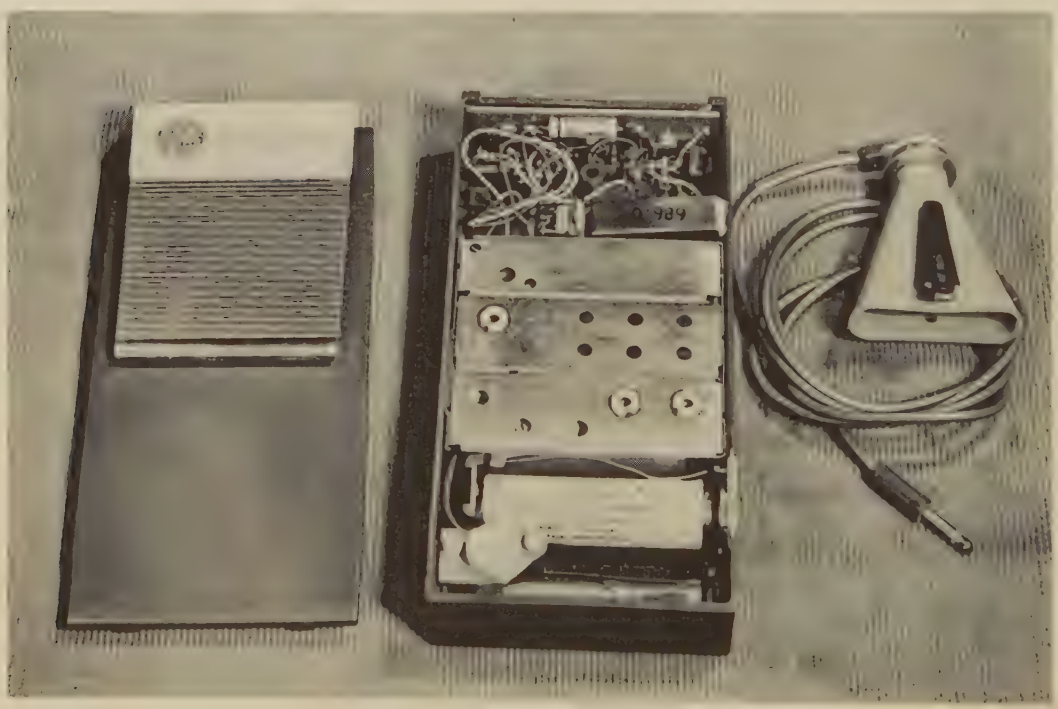


Fig. 5 - 150 Mc Personal FM Receiver - Interior with Signalling Option

INTRODUCTION TO PANEL DISCUSSION ON MOBILE RADIO VEHICULAR INTERFERENCE

John F. Chappell
U. S. Army Signal Research and Development Laboratory
Fort Monmouth, New Jersey

Vehicular electrical systems have the inherent capability of generating broad-band type impulsive interference, which, if effective control measures are not taken, will seriously degrade the performance of mobile communications receivers installed in the vehicle or installed in nearby vehicles. Thus we have two distinct problems: one, the interference to the communications receiver by the vehicular carrier; two, interference to the communications receiver from the electrical systems of nearby vehicles.

The term broad-band interference describes the spectrum character of this interference which manifests its presence over a broad spectrum of frequencies extending from the very low radio frequencies to at least 1,000 mc. Figure 1 illustrates some of the properties of broad-band impulsive type interference. It will be noted that the idealized case of a pulse of infinite height and infinitesimal duration will result in a uniform spectrum from dc to infinity. More realizable pulses of rectangular and triangular shape respectively would have a spectrum such as shown in the lower portion of Figure 1. It will be noted that the spectrum amplitude, in the flat portion of the spectrum is substantially a function of only the volt-time area of the pulse, whereas at the higher frequencies the spectrum amplitude is a function of both the volt-time area and the pulse shape.

As might be expected from the broad spectral energy distribution of an impulse, the energy accepted by, and the peak response of, a receiver to this spectrum are proportional to the receiver bandwidth. Also for bandwidths typical of mobile communications receivers there will be no significant overlapping of receiver responses due to the individual vehicular interference pulses. As a result of these facts a method of spectrum intensity measurement has evolved which has proven very useful in the measurement of vehicular interference. In this method which is illustrated in Figure 2 the spectrum intensity at the input of a receiver is expressed in terms of the r.m.s. sine-wave microvolts, per unit bandwidth of the receiver, which will produce a peak response in the receiver equal to that produced by the impulse spectrum.

The actual spectrum of vehicular interference would look something like that shown in Figure 3. The peaks and dips are brought about primarily by varying impedance to circulating currents on the vehicular body and wiring and the varying degree of radiation efficiency of the vehicular body at various frequencies. When it is considered that the r-f field generated by an unsuppressed vehicular electrical system can be of the order of volts per meter it is no wonder that we have an interference problem. Fortunately, however, a good receiver responds only to a small increment of the spectrum. For example, an ideal receiver would respond only to those frequency components falling within its pass band. Data taken on unsuppressed vehicles indicates interference signals of the order of 1,000 microvolts/meter for IF bandwidths of the order of 10 KC at a distance of 5 ft. from the vehicle. The suppression applied to the vehicle, therefore should be designed to reduce the interference spectrum level to a value such that the spectrum components falling within the receiver pass band will not be significant compared with the minimum usable desired signal.

An illustration of the receiver problem is shown in Figure 4. Here we have shown the possible points of entry of an interfering signal. The antenna paths, the path via non-signal leads, and case penetration through shield openings. With regard to the antenna path, the receiver may respond not only to the frequency components within its pass-band but, through spurious responses, may respond to frequency components outside the normal pass band. Also to be considered are intermodulation products generated by receiver non-linearities prior to the first mixer. In the case of the remaining two paths of signal entry, i.e. through shield leakage or conduction into the set on non-signal leads, the spurious energy can, once it gains entry into the receiver housing, couple to sensitive circuits and thereby degrade performance of the receiver.

I think it is very evident from the vehicular interference problem as just outlined that there are three groups with primary concern in this area. They are: (a) the user of the mobile communications equipment, (b) the mobile communications equipment vendor, and (c) the automotive vendor. The user is primarily concerned with satisfactory communications. In order to achieve this aim the communications receiver and the automotive vehicle must be compatible. To achieve compatibility the automotive vendor must

find effective and economical means of suppression which will reduce vehicular interference to acceptable levels for well designed mobile communications receivers. The mobile communications equipment vendor, on the other hand, should design his receiving equipment so that it is not unnecessarily susceptible to vehicular interference. It is readily evident that to bring about a successful marriage of the mobile communications receiver with the vehicular carrier it is highly desirable that we have standard test methods, together with appropriate test limits, that can be used to determine when the vehicular carrier has been suppressed to acceptable interference levels and to determine when the impulse interference susceptibility of the mobile communications receiver has been reduced to acceptable values.

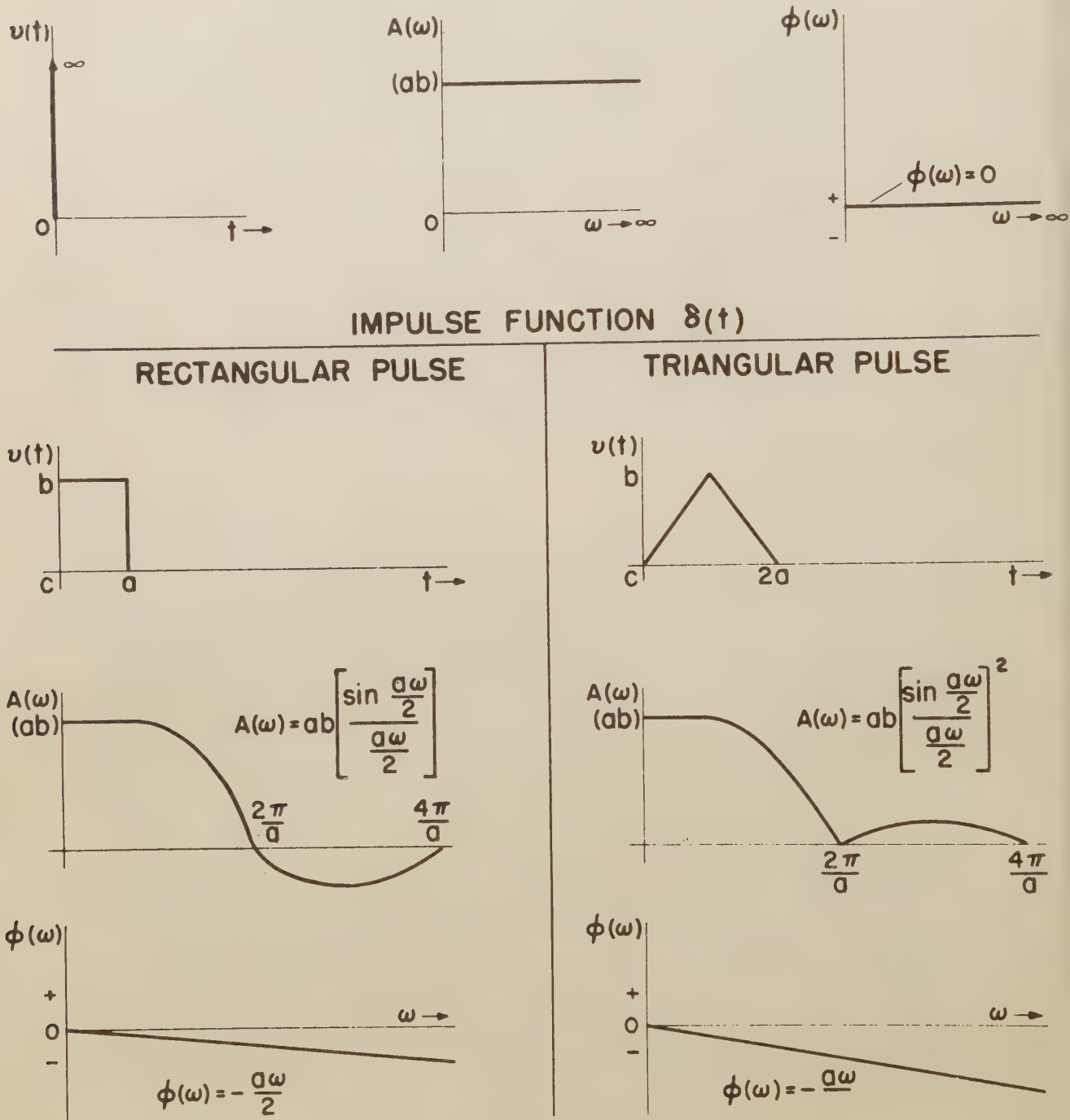


Fig. 1 - Spectrum Characteristics of Pulses

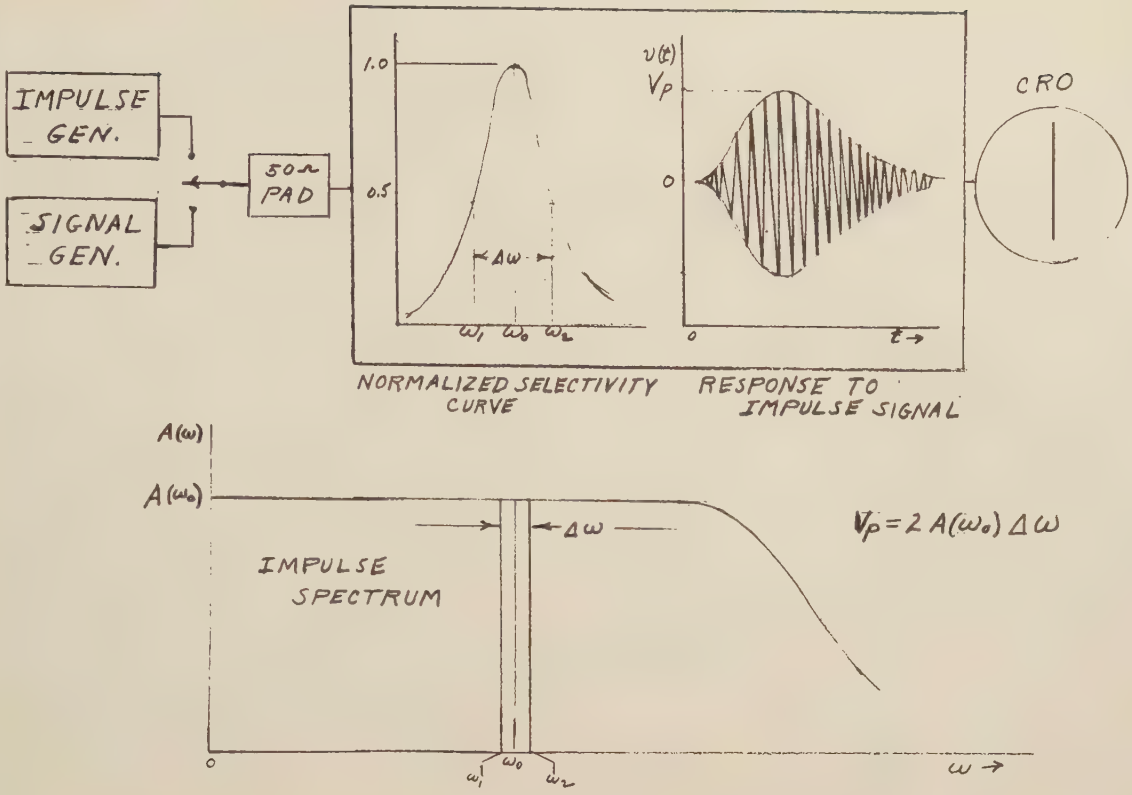


Fig. 2 - Method of Spectrum Measurement

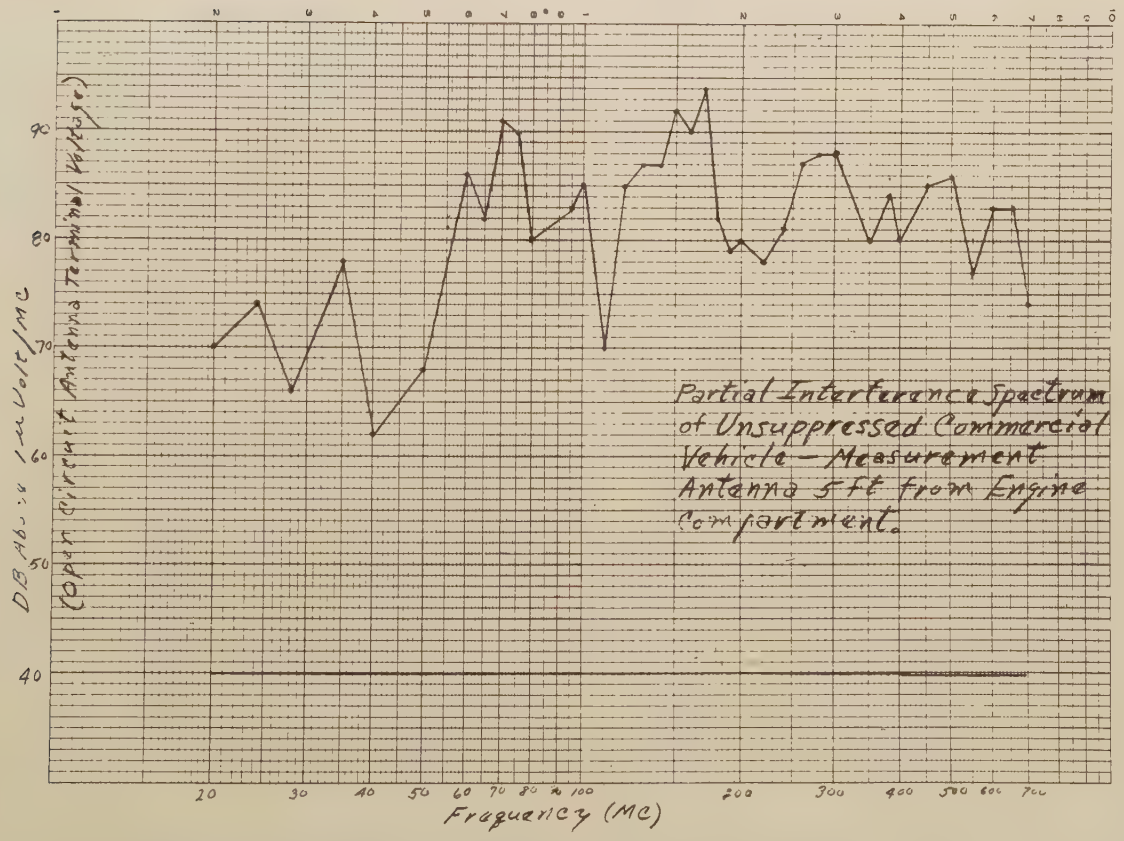


Fig. 3 - Typical Measured Spectrum of Vehicular Interference over Frequency Range 20 to 700 Mc

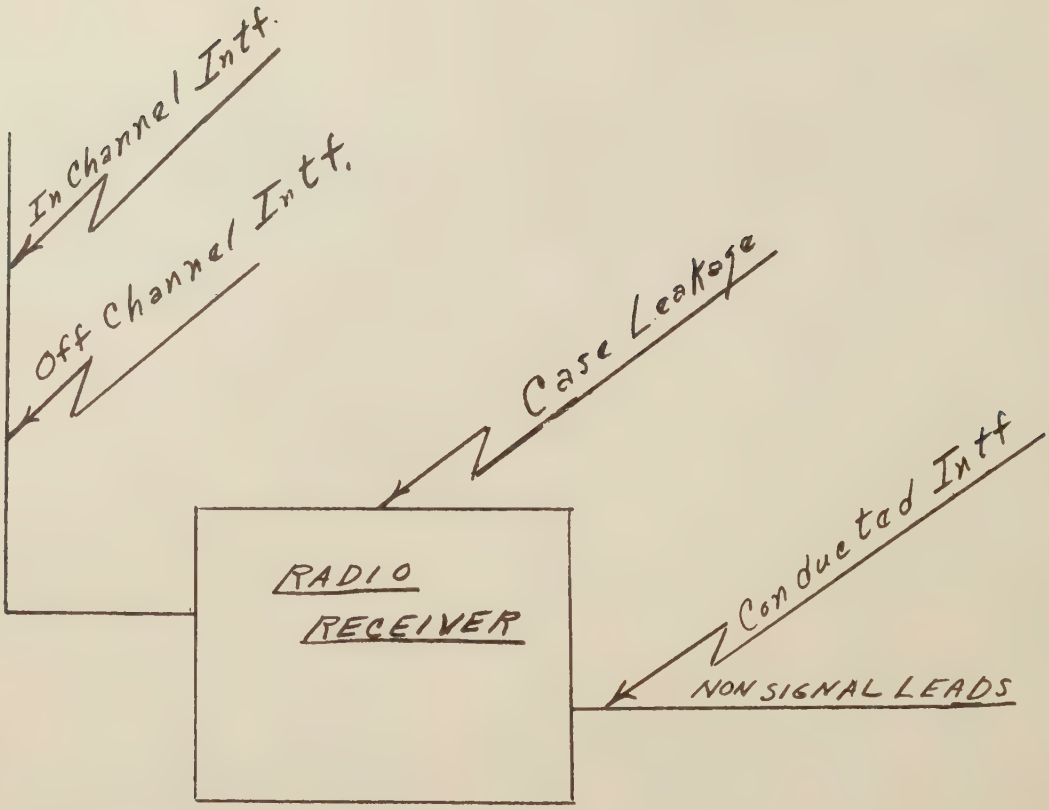


Fig. 4 - Possible Points of Entry of Vehicular Interference

PROPOSED PROCEDURES AND STANDARDS FOR MEASUREMENT OF VEHICULAR INTERFERENCE[1]

W. A. Shipman
Columbia Gas System Service Corporation
New York, N. Y.

My contact with interference, particularly mobile interference, dates back to 1946 when The Columbia Gas System Operating Companies first started mobile radio installations. Interference problems usually start with the first installation. For about eight or nine years, our companies had varying amounts of interference troubles, and we firmly believed that the thing to do to correct such conditions was to complain loud enough and long enough and the radio vendor would come up with some answers. Each year we had more mobile units, and exposure to interference increased. Channel loading was also increasing, and narrow band sets were getting closer and closer, potentially increasing interference.

In order to meet this problem, IRE formed Committee 27 - Radio Interference - in 1955, bringing together many people who had been working on interference problems with TV receivers, industrial machines, welders, induction heaters, and the like. I was asked to serve as temporary chairman of 27.7 - sub-committee for Mobile Radio Interference. Our first meetings were spent in surveying the mobile field. We decided that it would be extremely helpful if we had a standard method of measuring vehicular interference caused by the electrical system of the vehicle itself, such as ignition, generator, voltage regulator, and the like. The early committee consisted of representatives of the major mobile manufacturers, signal corps and user group.

It was recognized early in this effort that in order to postulate a useful standard method of mobile interference measurement, it would be necessary to enlist the aid and assistance of the automobile manufacturer's representatives. As a result, representatives of the automobile industry who were interested in interference problems and had been working with SAE Committees on TV interference caused by automotive vehicles were asked to join 27.7 efforts.

Committee 27.7 followed the TVI-SAE tests with great interest. It was thought possible by some, that, if the cars could be suppressed for TVI, this might go a long way towards solution of our two-way mobile interference problems. Early tests were, however, fragmentary, and there appeared to be no correlation between the TV interference at greater distances and the highly complex fields within 4 or 5 feet of the vehicle as well as interference field acting upon the VHF unit mounted in the vehicle.

As a result of these early tests, an approach independent of these SAE efforts was thought best, and an ad hoc committee of 27.7 was appointed to draft a proposed standard of measurement that would produce reproducible results and correlate to interference experiences, both for the VHF radio in the vehicle and the VHF unit nearby the vehicle. This group produced a draft of a standard, modeled after the Signal Corps research and specifications which was for presentation to the full 27.7 committee.

The auto manufacturer representatives on Committee 27.7 objected to the implied stringency of the drafted standard. Using suppression techniques suggested by the Signal Corps method might involve more suppression than was necessary and would involve additional expenses. They suggested the possibility that radio manufacturers could improve their equipment.

Time passed and work continued. Draft two followed draft one, and draft two was superseded by draft three. Both auto manufacturer representatives and radio manufacturer representatives making contributions and concessions. In the meantime, the radio manufacturers formed TR 8.10, an E.I.A. Committee, in order to study the draft of measurement standard and establish limits of acceptable interference for the equipment available commercially.

In November, 1957, a task force made

up of radio and automobile manufacturer representatives met at Anderson, Indiana, Proving Grounds of Delco-Remy, to determine (1) how much assistance the two-way mobile people would derive from the tentative limits established for TVI by SAE and, (2) what improvements could be expected in two-way radio by installing various other suppression components.

Similar tests were conducted at General Electric, DuMont, and RCA.

It might be well to describe briefly the method of measurement which draft three used. The Proposed Standard[2] describes a method for measuring the spectrum intensity of broad-band impulsive type interference. The spectrum intensity at each desired frequency is measured as the open circuit antenna terminal voltage, in terms of microvolts per megacycle bandwidth. The measurement is accomplished by a substitution method using a calibrated impulse generator which generates pulses of sufficiently short duration so as to result in a spectrum of frequency components of constant amplitude out to at least 500 MC, the highest frequency of measurement. The output of the calibrated impulse generator is series-injected into the antenna circuit of the receiving antenna in such a manner as to make the absolute value of the measurement independent of antenna and instrument input impedances. When the interference and the calibrated impulse generator inputs, applied consecutively, produce the same peak response from the interference measuring instrument, the interference spectrum intensity is equal to that of the impulse generator output minus the sum of the 20 db attenuation of the calibrated injection network and the loss occurring in the cable connecting the impulse generator of the injection network.

Readings are taken at two locations, (front and rear) in the frequency range of 25--500 MC. Interference peaks in between any two test frequencies which exceed either by 6 db, are measured and recorded. Vehicle and test equipment antenna arrangement will be shown in slides below. The measurement equipment is located 20' directly behind the measurement antenna. At each test frequency, the interference recorded is the maximum value observed when the vehicular engine speed is varied smoothly from idle to race.

In the meantime, draft number three was submitted to Committee 27 and returned to 27.7 with comments. It was also submitted to Committee 16 and other interested parties for comments.

Members of TR 8.10 and 27.7 met again with representatives of the automotive industry in Anderson on November 16-19. The modified near-field method (termed IRE Test) was used. Again SAE tests and IRE tests were made on the same vehicle. Degradation tests were also made on some of these same vehicles with low band antennas clamped to the bumper and high band antennas mounted on top of the vehicle with suction cups.

There was a mass of data taken at Anderson. Obviously, in the short time that has elapsed, the data has not been completely analyzed, although certain preliminary tries have been made at correlation. Of course, on a purely statistical basis, the data is fragmentary. One test on one car of a year's production is a very meager sample. The question of a closer simulation to actual installations with the degradation tests in place of IRE near-field tests is open to serious question. Degradation and IRE test data shows very poor correlation unless hand-picked. Very little better correlation is evident in comparison of IRE and SAE, adding further evidence to the belief that the near-field does not correlate well with the fields at greater distances.

This gives you a general idea where we have been and generally where we now stand. It is certainly evident that minimum incidental radiation in the 30 and 150 MC band does not appear to be a prerequisite to the design of electrical systems for vehicles offered as standard production. In fact, the "interference factor" (to coin a term) appears to be getting worse instead of better.

Were it not for the cost of test equipment, about $3,000 to $3,500, the user group could perform the IRE tests on vehicles in their fleet, and the problem could be approached with statistical techniques, using a large mass of data. It is hoped that some larger users may be

able to do this. Perhaps the most direct way would be for the radio manufacturers to specify limits based on the IRE Standards for their equipment and the specifications of the buyers of vehicles include these limits in their purchase orders. Under such specifications, the auto manufacturers can give some idea of the costs of such suppression. Assuming any statistical or average approach, this would not say that each individual car or truck will behave on such averages. There will still be a field for technical ingenuity in numerous cases.

At the present time, methods of reducing the noise radiation in existing electrical systems normally come under the heading of a patch-work or containment. Certain techniques, such as resistance cables or resistor plugs, reduce the noise a substantial amount, but other techniques create a condition similar to a leaking pipe with an infinite number of holes--each "hole" representing a discrete radio frequency. Every new capacitor or bond added at specific locations in the electrical system may close off many holes but also may serve to increase the pipeline pressure. (Resistors are, of course, an exception in that they do not normally increase pressure.)

Perhaps a better solution for noise reduction would be to establish noise specifications at the time new designs are initiated by the automotive industry.

[1]Presented as part of a panel discussion on December 4 at the 10th National Meeting of the Professional Group on Vehicular Communications at St. Petersburg, Florida.

[2]This is a brief description of the method. For a more complete discussion, see the latest draft of the Proposed Standard.

EFFECT OF VEHICULAR GENERATED INTERFERENCE ON DOUBLE-SIDEBAND AND SINGLE-SIDEBAND A.M. RECEIVERS

Kenneth E. H. Backman
Motorola, Inc.
Chicago, Illinois

Summary

As this paper was presented as part of a panel discussion, it is rather brief. The purpose, however, is to illustrate the manner in which impulsive type interference such as that generated by automobiles, neon lights, electric motors, etc., in the form of arc discharges, effects double-sideband and single-sideband A.M. receivers. Also, how this interference compares with that in F.M. receivers.

Introduction

Good afternoon gentlemen. I would like to put you a little at ease by saying that I am not going to talk very long. As has been pointed out by the moderator, I am to tell you about the effects of vehicular interference on double-sideband A.M. and single-sideband receivers. However, since most of you are not too interested in single-sideband, it is probably just as well that I be specific, be brief, and be seated.

Nature of the Problem

Mr. Meyer has told you of some of the effects of vehicular interference on F.M. receivers. Although the effects on A.M. and single-sideband receivers are not vastly different, there are some basic differences which I would like to discuss.

Impulse noise, as radiated by automotive ignition systems, is more noticeable in A.M. than in F.M. systems since F.M. systems limit the amplitude of the disturbance and are dependent only upon the frequency variations present whereas, of course, in A.M. the output is directly related to the R.F. input amplitude variations. The same is of course true of single-sideband as it is basically A.M. with one sideband suppressed.

The tremendous crowding of the spectrum has resulted in the natural step of channel splitting which has even further aggravated the problem. Close channel spacing requires narrow steep sided filters which are susceptible to ringing when excited by any form of impulse noise. This ringing results in a stretching of the interfering pulse, thereby blanking the receiver for a longer period of time. This is true in A.M., F.M. or single-sideband and for this reason, impulse noise from a vehicle obviously has a greater effect on narrow band systems. So it becomes apparent that the contributing factors to interference in communication receivers are moving in opposite directions to that desired. The higher compression ratios of the newer vehicles require higher firing voltages resulting in greater impulse radiation. Simultaneously, channel splitting has forced the use of narrow band receivers which are more susceptible to this impulse noise. The result—the growing ignition interference problem we are all confronted with.

There are numerous methods of combating the problem. One is to make the receiver insensitive to this interference. Although this is not impossible, no one has yet provided the complete answer. Another approach, and of course the most common now, is the use of various types of suppression devices. The other obvious approach would be to eliminate the noise at the source. As I have heard Mr. Stuart Meyer so aptly put it, "If the noise isn't there, you can't hear it." Of course this indicates removal of the engine from the vehicle, but this has raised serious objection by Mr. Short and the automobile industry.

How Serious Is The Problem?

Just what is the magnitude of the interference problem we are dealing with? Experiments have shown that if a 160 Mc. receiver is operated in a heavy-traffic or urban area, the ambient R.F. noise is frequently as high as 20 db. above the inherent receiver noise as measured at the receiver output. This means that if a given signal plus noise to noise ratio must be obtained under these conditions without altering the receiver design, it would be necessary to increase the power output of the transmitter 100 times. A power increase of this magnitude is obviously impractical. I have implied here that the source of noise is automobiles. Although the source is usually automobile ignition systems, any spark discharge such as high tension lines, neon lights, and brush commutators are also impulse noise generators. In fact, in fringe areas, thermal

noise and ignition noise are sometimes much alike and can frequently be mistaken for one another.

I'm sure that all of you realize that a great deal of effort has been expended in the direction of vehicular interference suppression. In the course of this work, it has been determined that signal levels of 200 microvolts have been produced by an automobile engine operating at an R.P.M. equivalent to 30 M.P.H. About 100 impulses per second larger than 50 microvolts and about 200 impulses per second larger than one microvolt were obtained. It is known that the energy contained in such a series of impulses is made up of many frequencies extending from a frequency equal to the repetition rate up to hundreds of megacycles. Most of this energy is radiated with varying degrees of efficiency, depending upon the characteristics of the spark plugs, the ignition wire, etc. However, the receiver and its antenna system has a given bandwidth and will, therefore, accept those frequency components that fall within the selectivity of its front end. This energy so to speak, excites or rings the antenna tuned circuit. As the disturbance progresses to more selective circuits and narrower bandwidths, the ring becomes longer. To reiterate, the impulse response amplitude varies directly with the receiver bandwidth, while the length of time that the impulse response occurs varies inversely with the receiver bandwidth. Therefore, the audio impulse noise for a narrow band system will be less intense but will be present for a greater period of time as compared to a wide band system. Also, the squaring-off of the frequency response in order to obtain a steep response slope will cause a higher amplitude impulse response ringing which will last for a longer time than for a less rectangular frequency response.

Unfortunately as the frequency spectrum becomes more congested, tuned filters will have a more rectangular frequency response curve shape. This of course infers that impulse noise will become even more destructive to communications.

As I have already mentioned, impulse noise usually occurs in bursts. Because of this repetitive nature, it has a greater effect on A.M. or single-sideband systems. Let me explain. An F.M. receiver normally has fixed time constants which permit rapid recovery from a noise burst attack. A.M. or single-sideband usually employs A.G.C. which has a fast attack--slow decay characteristic. Therefore, in the presence of noise bursts, the A.G.C. will cause the gain of the single-sideband receiver to be low. Any small noise, i.e., noise lower than the carrier level, will ride on the carrier and set the A.G.C. level. In F.M., these low level noises of course have no effect as they are limited out. Similarly, changes in amplitude of the interfering signal above the carrier level would have no effect on F.M., whereas it would have a significant effect on A.M. or single-sideband. One factor in favor of the A.M. receiver, however is that amplitude limiting in the detector would tend to keep the interfering peaks down.

Conclusion

I have tried to point out some of the basic differences of the effects of noise interference on F.M. and A.M. or single-sideband systems. I believe that what I have said can be simply summed up by saying that because of the amplitude response nature of a single-sideband receiver, it is going to be inherently more susceptible to impulse noise interference and consequently, noisier.

VEHICULAR RADIO SUPPRESSION

Brooks H. Short, Director
Advance Engineering Department
Delco-Remy Division, General Motors Corporation
Anderson, Indiana

The present status of the automotive industry supported radio suppression project is covered. The international aspects of vehicular interference are also covered with an up to the minute report on the possibilities of international standardization. The cooperative program between the Electronic Industry Association, the Institute of Radio Engineers, and the Society of Automotive Engineers is discussed. This cooperative work must lead to the most economical overall installations. The final section of this paper describes new suppression devices that should materially help in the suppression of undesirable radio radiations.

Mr. Moderator, members of the panel, ladies and gentlemen: It is a pleasure to discuss this subject with you as a member of this panel. When one thinks of a panel one normally thinks of a group of people involving at least one pretty girl, as I look at this group I find we are sadly lacking in that area. Nevertheless, we shall attempt to cover this subject in spite of this limitation.

Undoubtedly you are interested in the present status of our Society of Automotive Engineers suppression program. We find that all of the passenger cars being manufactured in the United States can be suppressed to the SAE level by using high resistance type, high tension cables or by using resistor plugs and suppressed caps and rotors. When we determined this we thought our job was done and we recommended the adoption of our specification by the Automobile Manufacturers Association. When the proposal was being considered by the Automobile Manufacturers Association, it became apparent that our job was only partially done since the truck and heavy duty vehicle field have decided that they can not use suppressor type cables because of their poor reliability in this type of application. It became apparent at this time that a new group of suppression components would have to be developed and evaluated before our program could go into effect. Our various Engineering Departments are working in this area and we hope to have satisfactory solutions to the various problems in the near future. The SAE suppression specification can not be enforced across the board until all of these problems are solved.

The automobile industry is being faced with another suppression problem at this time. This problem is international in scope. Nearly every country in the world either has in effect or expects to adopt a suppression specification in the very near future. These specifications vary over wide limits. In some the measuring antenna is placed 50 feet from the vehicle, in others it is as close as two feet. In some countries horizontal polarization is used as a measuring criteria, in others vertical polarization is used. In still others the maximum is read irrespective of polarization. Some countries read in terms of "quasi" peak, others make measurements in terms of true peak. There are so many different possible combinations of the measurement parameters that it is very difficult to arrive at a given set of measurements that would cover all of the different specifications. The problem is further complicated by the different limits that are used and the different units of measurement. A Working Group has been set up by the CISPR to study the different specifications and attempt to arrive at one compromise specification that will most nearly meet all of the present specifications. A meeting of this group is expected early in 1960. I have been designated as a delegate to this working party from the United States of America. When a compromise arrangement has been arrived at, the automobile industry will attempt to provide suppression means for their products. In an attempt to get a common understanding of the problems, a meeting was held in Anderson, Indiana on June 22. This meeting was attended by representatives from Germany, Great Britain and the United States. At this meeting it was definitely determined that the British, German, and American instruments gave the same indication. This group also had a meeting of the minds of what parts of

the specifications could be reconciled.

All of the above may be academic as far as the two-way installed radio is concerned. Your various committees have been making remarkable headway in their work toward standardization of measurement in this field. We have reviewed the proposed specification as issued by the IRE Committee 27.7. We know that the EIA Committee TR 8.10 is in the process of determining satisfactory levels as read according to the 27.7 specification. When they have arrived at satisfactory limits the automobile industry can go to work to develop suppression techniques that will meet the requirements of your specification. In this connection I should like to reemphasize a point I made during your meeting in Chicago last year. That point is that it is our combined responsibility to arrive at the most economical approach for a given communication effectiveness. We all appreciate that suppression levels could be set at such a low value that one could install any wide open radio and get satisfactory communication. We all know that there are certain steps that can be taken in the design of the radio gear that will render receivers insensitive to interference. It appears desirable to weigh the cost of these devices against the cost of the custom made suppression so that we come out with the most realistic overall system.

I must point out that if the specification arrived at is extremely rigid complete shielding of the ignition system will be required. In many cars complete shielding would seriously interfere with efficient operation of the ignition system. This could prevent the car from running. Now I ask, what good would communications be if in order to get them through, the car would not run? For that reason we of the automobile industry feel that the various committees should take a realistic view and specify suppression limits that will not unfavorably effect the operation of the vehicle.

There are some new suppression techniques that should be very interesting to us. The first of these is a ring of ferrite material that was developed by the Phillips Company in Holland. A ring of this material when placed over the high tension lead causes a portion of that lead to act as a relatively high "Q" inductance, giving the effect of an inductance suppressor. We expect to investigate this new item in an effort to evaluate its performance as compared to a resistive suppressor element. A new type coaxial condenser has been developed in the United States that is very interesting. In this new condenser one foil is made heavy enough to carry the current that is normally conducted through the circuit. When so constructed the low voltage DC is fed at the inside of the condenser to one end of this foil. The outside end of the corresponding foil is connected to the outside circuit. The other foil is formed in the normal way and is grounded by low inductance approaches. In this way the capacitor acts as an infinite number of sections of "T" type filer. The attenuation characteristics of this type condenser are said to be infinitely better than the standard coaxial or feed through types.